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(54) **Method for calculating a weighting vector for an antenna array**

(57) The present invention generally relates to the field of wireless communication systems with high-speed mobile access, especially to pilot-assisted wireless communication systems considering channel estimation, and, more particularly, to a simple weighting vector calculation method to support interference cancellation in said multi-carrier systems, wherein Orthogonal Frequency Division Multiplexing (OFDM) is applied as a multi-carrier modulation technique.

In order to increase the carrier-to-interference-plus-noise ratio (CINR), a spatial filtering (beamforming) algorithm is applied, in which antenna weighting vectors $\underline{w}_k$ are calculated for each subcarrier k in such a way that the denominator of the associated Rayleigh quotient Γ_k, calculated by means of the autocorrelation matrix of the signal channel vectors $\underline{a}_k$, the interference channel vectors $\underline{b}_k$, said antenna weighting vectors $\underline{w}_k$ and the noise power $\sigma_K^2 \cdot I$ of the mobile radio channel, becomes a minimum, thereby maximizing the numerator of said Rayleigh quotient Γ_k. To perform a single interference cancellation in said calculation of the antenna weighting vectors $\underline{w}_k$, only one inner product operation and one vector subtraction is needed.

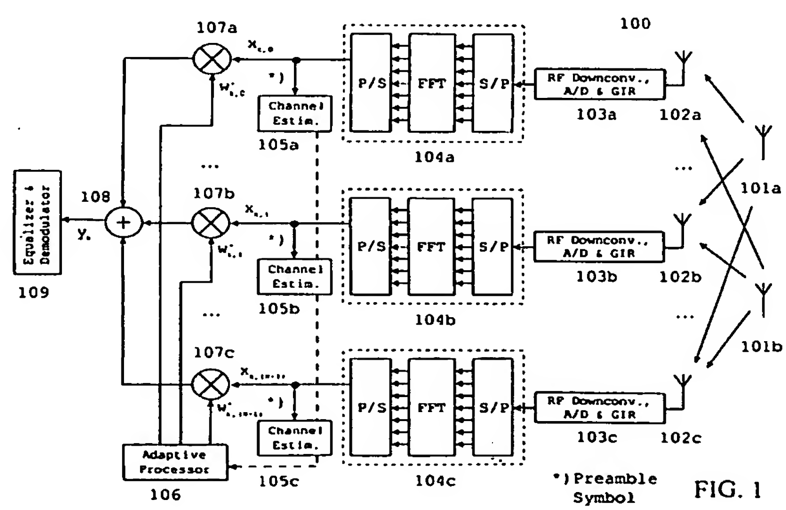

FIG. 1

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Description

FIELD AND BACKGROUND OF THE INVENTION

[0001] The present invention generally relates to the field of wireless communication systems with high-speed mobile access, especially to pilot-assisted wireless communication systems considering channel estimation, and, more particularly, to spatial filtering (beamforming) algorithms comprising a novel weighting vector calculation method to support interference cancellation in high-speed wireless multi-carrier systems, wherein Orthogonal Frequency Division Multiplexing (OFDM) is applied as a multi-carrier modulation technique.

[0002] Nowadays, the growing demand for mobile communications is constantly increasing the need for an enhancement of its capacity and quality of service (QoS). Third and fourth generation cellular systems must be designed to handle the demand for higher data rates and quality in wireless communication systems in a cost-effective and flexible manner. To achieve these objects, a mobile transmission system must be able to cope with a plurality of impairments.

[0003] On the one hand a broadband radio channel, as needed for the transmission of high data rates, is characterized by severe attenuation fades (frequency-selective fading) caused by multipath propagation of the transmitted mobile radio signals. The wireless environment is typically characterized by a plurality of scattering processes and the absence of direct line of sight (LOS) paths between the base stations (BS) and the mobile stations (MS). On the other hand the channel exhibits a time-variant behavior due to the mobility of the receiver, which possibly requires a continuous adaptation of the transmission system to said behavior. For example, in TDMA systems the mobile radio channel causes perturbation effects on the transmission link in the form of attenuation, distortions and fluctuations.

[0004] Adaptive antennas promise to greatly increase the capacity of cellular systems by suppressing co-channel interference, improving coverage quality, and mitigating multipath interference. In the past decade, much research has been done towards merging adaptive antenna technology with existing cellular and PCS air-interface specifications based on AMPS, TDMA and CDMA technology as well as OFDM. Some of the inherent advantages of OFDM are its flexibility, scalability and the ability to communicate a wideband signal without using temporal equalization schemes.

Space diversity and antenna arrays

[0005] Space diversity is a way to considerably improve the performance of a mobile transmission system. For this purpose, antenna arrays can be used to combat various types of channel impairments. An antenna array consists of a set of antennas designed to receive signals radiating from specific directions and to attenuate signals radiating from other directions of no interest. The output of said array elements is weighted and added by means of a so-called beamformer to produce a directed main beam and adjustable nulls. In order to reject the interferences, the beamformer has to place its nulls in the direction of possible interferers, and to steer to the direction of the target signal by maintaining constant gain at this direction.

[0006] An antenna array with sufficient antenna spacing can provide spatial diversity to mitigate multipath fading. For this reason the distance between the antennas must be large enough to ensure that signals received by different antennas are uncorrelated: a spacing of at least half a wavelength is required. A distinction must be made between the following two types: micro-diversity, where the distance between different antennas is in the order of only a few wavelenghts, and macro-diversity, where said distance is at least 10 wavelenghts. In practice, this usually means that in case of micro-diversity all antennas are located at the same station, and in case of macro-diversity all antennas are located at physically separated stations. A combination of both types is often used to combat both effects caused by multipath fading (micro-diversity) and by shadowing (macro-diversity).

[0007] If the antenna elements are adequately spaced, the receiver signals can be considered as being independent, and thereby the probability that all received signals fade at the same time can drastically be reduced. Beamforming and diversity reception can be employed to cope with the effect of delay spread and co-channel interference (CCI). Consequently, the performance of the entire mobile transmission system is improved. The advances in technology and the expanding demands for antenna arrays have made them very economical, and the trend of using GHz carriers for wireless access networks has significantly reduced the dimensional requirements of said antenna arrays.

[0008] The objective is to apply a set of complex weights to the antennas that produces signal reinforcement at the mobile receiver. Early descriptions of adaptive transmitter diversity dealt only with single-user systems. More recently, uplink measurements have been used to select antenna weights that support a plurality of simultaneous users by nulling or mimimizing the interference experienced at each mobile station (MS) from transmissions intended for other mobile stations.

[0009] Today it is common practice to use diversity arrays at a base station (BS) receiver to mitigate multipath fading and to support multiple users. Recent approaches to transmitter diversity distinguish the different antennas by relative delays, different spreading codes or by space-time codes, none of which requires or uses information on the current channel state. However, if such information is available from uplink measurements that are nearby in time and fre-

quency, then adaptive transmitter diversity methods with potentially better performance can be developed.

[0010] The performance of transmission systems employing antenna arrays can be measured by means of the signal-to-noise ratio (SNR) and mutual information. Although the two metrics are closely related, they have important differences: The SNR characterizes the performance of typical uncoded systems, whereas mutual information measures determine the maximum rate of reliable communication achievable with coded systems (in the absence of delay and processing constraints).

[0011] When the message is intended for a single recipient, a beamforming strategy is optimal. With beamforming, the transmissions from the different antenna elements at the base are designed to be coherently added at the intended receiver, yielding an average enhancement of the SNR and a corresponding enhancement of the mutual information over single-element antenna systems. However, this improvement requires that the transmitter antenna array has accurate knowledge of the channel parameters to the intended recipient, which is difficult to achieve when the parameters are time-varying. Gains obtained in practice with only partial information at the transmitter are more modest as a result. In addition, in broadcast scenarios this factor of enhancement cannot be obtained at each receiver even when the parameters of all channels are perfectly known as it is generally not possible to simultaneously beamform to multiple recipients.

[0012] Antenna arrays may be employed either at the transmitter or the receiver. In a mobile radio system, it is generally most practical to employ an antenna array at the base station rather than at the mobile units. Then, for transmitting from the mobile station (MS) to the base station (BS) during the uplink, diversity is achieved by means of multiple-element receive antenna arrays ("receiver diversity"), whereas in transmitting from the base station (BS) to the mobile stations (MS) during the downlink, diversity is achieved by means of a multiple-element transmit antenna array ("transmitter diversity"). Transmitter diversity has traditionally been viewed as more difficult to exploit than receiver diversity, in part because the transmitter is assumed to know less about the channel than the receiver, and in part because of the challenging signal design problem: the transmitter is permitted to generate different signals at each antenna element.

[0013] For a receiver-adaptive antenna array, training or pilot signals, along with the internal structure of the message-carrying signals, enable the receiver to increase its signal-to-noise ratio (SNR) and carrier-to-interference ratio (CIR). For a transmitter-adaptive antenna array, usually either a-priori or auxiliary information concerning the propagation of signals from the antenna array to the desired (in-cell) and undesired (out-of-cell) receivers is needed.

[0014] Space-time processing promises to significantly improve mobile radio performance. The principles of space-time processing can be used to develop "smart" antennas that employ adaptive arrays of antenna sensors. Therefore, the "smart antenna" concept has become very interesting for the mobile communications industry. The term "smart antenna" usually refers to the deployment of multiple antennas at the base station (BS), coupled with special processing of the multiple received signals. Smart antennas can adaptively reject co-channel interference (CCI) and mitigate effects caused by multipath fading. They have been identified as a promising means to extend the base-station coverage, increase the system capacity and enhance the quality of servive (QoS).

[0015] A "smart" antenna generally consists of the sensor array, the beamforming network and the adaptive processor.

- Sensor array: To receive (and transmit) signals, the sensor array comprises N antenna elements (sensors). The physical arrangement of the array (linear, circular, etc.) is arbitrary, but places fundamental limitations on the capability of the "smart" antenna.

- Beamforming network: The output of each of these N antenna elements is fed into the beamforming network, where the outputs are processed by means of linear time-variant (LTV) filters. These filters determine the directional pattern of the "smart" antenna, that means the relative sensibility of response to signals for a specified frequency from various directions. The outputs of the LTV filters are then summed to form the overall output y(t). The complex weights of the LTV filters are determined by the adaptive processor.

- Adaptive processor: As mentioned above, the adaptive processor determines the complex weights of the beamforming network. The signals and known system features used to compute said weights include the following items: the signals $x_i(n)$ (for $i = 0,...,N-1$) received by the antenna array, the output y(n) of the "smart" antenna, the spatial structure of the antenna array, the temporal structure of the received signal, feedback signals from the mobile stations (MS), and the network topology.

[0016] For the general case, where the receiver employs an array of antenna elements, a vector channel model is used. Therein, the channel model contains both the temporal and spatial characteristics of the channel.

BRIEF DESCRIPTION OF THE PRESENT STATE OF THE ART

[0017] According to the present state of the art, there are different solutions available to the problem of adaptive antenna processing concerning techniques for performing an interference cancellation, each of them being optimized to a specific environment. Thus, each of these solutions inherently contains certain restrictions. In order to representatively explain some of the most important solutions, it is necessary to briefly describe their main aspects.

Power control and beamforming

[0018] As mentioned above, one of the main impairments that degrades the performance of a wireless link in a cellular radio system is co-channel interference (CCI). Two important approaches for improving the performance in wireless networks by appropriately controlling the co-channel interference are power control and spatial filtering (antenna beamforming).

[0019] In power control, the transmitter powers are constantly adjusted by increasing if the signal-to-noise ratio (SNR) is low, and decreasing if the SNR is high, such that the quality of weak communication links can be improved. Receivers employing antenna arrays may adjust their beam patterns in such a way that they obtain maximum gain towards the directions of their transmitters and minimum gain towards the other directions, such that the overall interference power can be minimized.

[0020] Linear processing using antenna arrays is a signal processing technique called spatial filtering or "beamforming". A beamformer uses an array of antenna elements to exploit the spatial separation of impinging signals that mutually overlap in their spectra. This spatial filtering process enables the beamformer to separate the desired signals from the interferering signals by adaptively updating its weights and steering beams towards the desired users.

[0021] Thus, the purpose of an adaptive beamforming in the downlink is to modify the beam pattern in order to enhance the reception of the desired signal at the mobile station (MS), while simultaneously suppressing interfering signals by means of a complex weight selection. In other words, beams are formed in such a way that they point in the direction of the desired user in order to minimize the energy transmitted in direction of interfering sources. By properly adjusting the phase of each antenna, the main lobe of the antenna pattern can be directed to the desired angle. This enhances the strength of the desired signal and also suppresses the interference from signals coming from undesired directions. Additionally, adaptive beamforming reduces multipath fading of the transmitted signals by using narrow beams.

[0022] Many weight adaptation algorithms are known from literature. They all combine signals from multiple antennas in order to satisfy specific optimization criteria. These criteria comprise methods for minimizing the mean square error (MSE), maximizing the signal-to-noise ratio (SNR), and maximizing variance. These algorithms find applications in both spatial processing and temporal processing. The three most commonly used adaptation techniques are: Least Mean Squares (LMS), Direct Matrix Inversion (DMI) and Recursive Least Squares (RLS). In the following, two commonly used beamforming algorithms shall representatively be described.

[0023] Common adaptation techniques used in spatial and temporal processing

[0024] The main goal of common adaptive algorithms used in spatial and temporal processing is to maximize the output signal-to-interference-plus-noise ratio (SINR). This is accomplished by minimizing the cost functions associated with various criteria, e.g. Minimum Mean Square Error (MMSE) and Least Squares (LS) criteria.

[0025] In the following, three algorithms for adaptively updating a finite impulse response (FIR) filter - spatial or temporal - shall briefly be described. These techniques can also easily be extended to perform channel estimation. Thereby, the signal estimate y(n) is assumed to be estimated from a finite collection of received data samples $\underline{x}(n)$ with an FIR filter or weight vector $\underline{w}$. This can be expressed with the following inner product:

$$y(n) = \underline{w}^H \underline{x}(n),$$

wherein

N	is the number of antenna elements,
$\underline{w} \in C^N$	denotes the complex-valued weighting vector,
$\underline{x}(n) \in C^N$	denotes the complex-valued received signal vector,
$\underline{y}(n) \in C$	denotes the complex-valued output signal, and
'H'	denotes the complex conjugate transpose operation (Hermitian transpose).

[0026] The vector $\underline{x}(n)$ can be a collection of past time samples (in the case of temporal filtering), or a collection of antenna outputs (in the case of spatial filtering).

1. Minimum Mean Square Error (MMSE) algorithm

[0027] The cost function $J(\underline{w})$ for the MMSE criterion is the expected value of the square error between the beamformer (or equalizer) output $y(n)$ and the desired version $d(n)$ of that signal:

$$J(\underline{w}) := E\{|y(n)-d(n)|^2\}.$$

[0028] The optimal solution for the weights can be found by means of the following equation:

$$\underline{w}^{opt} = \underline{\underline{A}}^{-1}\,\underline{p},$$

wherein

$\underline{\underline{A}} := E\{\underline{x}(n)\underline{x}(n)^H\} \in C^{N\times N}$	is the complex-valued space-time correlation matrix of the time-variant sensor data $\underline{x}(n) \in C^N$ from N antenna elements with the discrete time variable n,
$\underline{p} := E\{\underline{x}(n)\cdot d(n)^*\} \in C^N$	is the correlation of the complex-valued sensor data $\underline{x}(n) \in C^N$ with the desired signal d (n) $\in$ C,
'*'	denotes the complex conjugate operation, and
'H'	is the complex conjugate transpose operation (Hermitian transpose).

2. Least Mean Squares (LMS) algorithm

[0029] Another algorithm which is frequently used in antenna array processing is the time-domain Least Mean Square (LMS) adaptive algorithm. The LMS algorithm uses a minimum mean square error criterion to determine the appropriate antenna weighting vectors $\underline{w}$. This algorithm is considered as an optimal algorithm because the solution minimizes the error between the array output and the desired signal. Therefore, it is assumed that the desired signal is known, or a signal containing the desired signal characteristics is available. The LMS iterative equations are:

$$\underline{w}(n+1) = \underline{w}(n) + \mu\cdot\underline{x}(n)^*\cdot\varepsilon(n)\ \forall n,$$

wherein

μ	is the constant stepsize, which governs the rate of convergence of said iterative process,
$\varepsilon(n) = d(n)-y(n)$	is the error function between the desired signal d(n) and the output y(n), and
'*'	denotes the complex conjugate operation.

[0030] LMS is a computationally simple algorithm. Its complexity grows linearly with the number of antenna elements N. However, it suffers from slow convergence. Unlike the MMSE criterion, the LS criterion tries to minimize the time-average error between the linear processor output and a desired response over a finite number of time samples. The cost function $J(\underline{w})$ for the LS criterion is:

$$J(\underline{w}) = \left|\sum_{i=0}^{N-1}\underline{w}^H\underline{x}_i - d_i(n)\right|^2,$$

wherein

$\underline{x}_i \in C^N$	is the i-th received data vector,
$d_i(n) \in C$	is the i-th desired signal at the discrete time n,
N	is the number of the antenna elements, and
$\underline{w}^H\underline{x}_i$	denotes the inner product of the vectors $\underline{w}$ and $\underline{x}_i$.

[0031] The optimal weight vector that forces the LS gradient function to zero is given by

$$\underline{w} = (\underline{\underline{X}}^H \underline{\underline{X}})^{-1} \underline{\underline{X}}^H \underline{d}(n) \in C^N,$$

wherein

$\underline{\underline{X}} := [\underline{x}_0, \ldots, x_{N-1}] \in C^{N \times N}$	is the received data matrix,
$\underline{d}(n) := [d_0(n), \ldots, d_{N-1}(n)]^T \in C^N$	denotes the desired signal vector, and
'H'	is the complex conjugate transpose operation (Hermitian transpose).

3. Maximum power beamforming algorithm

[0032] Recently, eigenvector-based methods have been developed to perform a maximum power beamforming for randomly spaced antenna array systems. These algorithms use the autocorrelation matrix $\underline{\underline{A}}$ of the sensor data to find the weighting filters that pick out the signal with the highest peak power spectral density. Thereby, antenna weighting vectors $\underline{w}$ are chosen to solve the following maximization problem,

$$\underline{w}^{opt} = \arg \max_{\underline{w}} \left\{ \underline{w}^H \underline{\underline{A}} \underline{w} \right\}, \text{ subject to } \|\underline{w}\|_2 = 1,$$

wherein

$\underline{\underline{A}} := E\{\underline{x}(n)\underline{x}(n)^H\} \in C^{N \times N}$	is the complex-valued space-time correlation matrix of the time-variant sensor data $\underline{x}(n) \in C^N$ from N antenna elements with the discrete time variable n,
$\underline{w} \in C^N$	is the desired complex-valued weighting vector,
$\underline{w}^H$	denotes the complex conjugate transpose operation (Hermitian transpose) of $\underline{w}$,
$\|\underline{w}\|_2 := \sqrt{\underline{w}^H \underline{w}}$	denotes the Euclidian length of $\underline{w}$,
$\underline{w}^H \underline{w} := \sum_{i=0}^{N-1} \|w_i\|^2$	is the inner product of $\underline{w}$ with $\underline{w}$.

[0033] The desired weighting vector $\underline{w}$ is given by the eigenvector corresponding to the largest eigenvalue of the autocorrelation matrix $\underline{\underline{A}}$. The bulk of the computation involved in maximum power beamforming is involved in the following steps:

a)calculating the autocorrelation matrix $\underline{\underline{A}}$ from the sensor data $\underline{x}(n)$, and

b)performing the eigenvector decomposition of the autocorrelation matrix $\underline{\underline{A}}$.

[0034] The power method of eigenvector decomposition provides a low-computation, iterative method to find the eigenvector with the largest eigenvalue.

Single-channel signal extraction algorithms

[0035] Common single-channel signal extraction algorithms include interference rejection and joint detection (JD) techniques. In all these algorithms, only temporal processing is utilized since the receiver antenna at the base station (BS) contains only one element.

[0036] Interference rejection algorithms can be divided into non-blind and blind interference rejection techniques that only employ temporal processing. The non-blind techniques employ some sort of training to estimate channel and receiver parameters with adaptive processing. On the contrary, in blind techniques training sequences are not available and channel is estimated using some known structure of the signals. In the following, only non-blind interference rejection techniques shall be considered.

[0037] Non-blind adaptive processing techniques can be broken into three main approaches: Linear Time-Independent Adaptive Filtering (LTIAF), Linear Time-Dependent Adaptive Filtering (LTDAF), and non-linear adaptive processing using Decision Feedback Equalizers (DFE) or Maximum Likelihood Sequence Estimators (MLSE). In order to give a brief overview, some of the most frequently used techniques shall representatively be described.

[0038] Despite their low computational complexities and simple structures, Minimum Mean Square Error (MMSE) linear equalizers - LTIAF and LTDAF - are not very efficient on channels with deep spectral nulls in the passband. This is because the linear equalizer places high gain near the spectral null in order to compensate for the distortion, thereby enhancing the noise present in those frequencies. Non-linear methods do not suffer from this phenomenon. One of the most common forms of a non-linear adaptive processor is the Decision Feedback Equalizer (DFE), which is commonly used as an equalizer to mitigate the intersymbol interference (ISI) in the channel. In addition to ISI mitigation, DFE can perform limited interference rejection. In the literature adaptive, fractionally spaced DFE have been proposed to cancel co-channel interference (CCI) and suppress ISI in the presence of a single, dominant co-channel signal and uncorrelated, additive Gaussian noise.

[0039] The above-mentioned interference rejection techniques attempt to estimate interfering signals and then strip them from the total signal, leaving only the desired signal components plus noise. On the other hand joint detection (JD) algorithms recover all the signals, desired and interfering, from the signal environment, and then discard the latter. Said algorithms are based on the Maximum Likelihood (ML) and Maximum a Posteriori (MAP) criteria for the joint recovery of the co-channel signals. These criteria are used to derive two important sequence estimation and symbol-by-symbol detection techniques, Maximum Likelihood Sequence Estimation (MLSE) and Maximum A Posteriori Symbol Detection (MAPSD), respectively.

Multi-channel signal extraction algorithms

[0040] In a mobile communication system, effects caused by multipath propagation of the signals to be transmitted and co-channel interference (CCI) are the major impairments to the signal quality and system capacity. Multipath propagation gives rise to fading and time dispersion. The time dispersion problem can be solved using linear equalizers or non-linear equalization techniques such as Decision Feedback Equalization (DFE) and Maximum Likelihood Sequence Estimation (MLSE). Multipath fading can be mitigated by antenna diversity at the receiver.

[0041] These multiple antennas collect more signal energy and diversity gain due to the spatial separation of their antenna elements. When the antennas are spaced appropriately, there is a good chance that not all of them will fade at the same time. Moreover, multiple antennas can be used to combine multiple copies of both desired and interfering signals in such a way that the desired signal components can constructively be added, whereas the interfering signals add destructively. This process of exploiting spatial diversity is called spatial equalization. Space-Time Adaptive Processing (STAP) receivers combine spatial and temporal equalization in order to provide better interference rejection performance as well as better intersymbol interference (ISI) reduction than single antenna receivers. Similar to single antenna array processing techniques, STAP algorithms can be classified according to how they treat interference: interference rejection or joint detection.

[0042] Non-blind interference rejection techniques require the use of training sequences to estimate the mobile radio channel. Although the use of training sequences greatly simplifies the channel estimation problem, exploiting them can be difficult when interfering users transmit asynchronously. Non-blind interference rejection techniques can be broken into linear, non-linear, and hybrid categories. Linear interference rejection tends to break down in overloaded environments. By contrast, hybrid techniques combining linear and non-linear approaches have been proposed that try to combine the advantages of each approach.

[0043] In the literature, linear signal extraction techniques can be classified as spatio-temporal techniques. Thereby, the antenna weights are optimized using the MMSE criterion. Their object is to maximize the output signal-to-interference-plus-noise ratio (SINR).

[0044] Adaptive beamforming algorithms for signal acquisition in digital radio systems are known from the prior art. Linear beamformer are used for steering nulls in the direction of the interfering signals. Additionally, a frame synchronization is achieved by locating the peaks in the cross correlation of the beamformer outputs with modified training sequences. The beamformer weights are updated using a LS criterion, and an equalization of intersymbol interference (ISI) is carried out using a fractionally spaced linear equalizer. The algorithm can effectively separate several users. However, its performance is limited by the linear beamformer, that means as long as the number of co-channel users does not exceed the number of antenna elements, a linear beamformer (N-element array) can null out up to N-1 users.

[0045] Many signal extraction and interference rejection algorithms concatenate linear array processing with non-linear temporal processing such as DFE and MLSE. In general, non-linear adaptive array processing techniques perform much better than the aforementioned linear techniques, especially in severe multipath fading environments.

[0046] Joint detection (JD) receivers are capable of eliminating intracell interference, that means the interference caused by other users in the same time slot and cell, with a reasonable effort and complexity. The purpose of joint

space-time processing is to obtain spatial autocorrelation matrices which can directly be used for the calculation of the beamforming weights needed for the downlink. The spatial autocorrelation matrix of the interference $\underline{\underline{B}}$ is computed by estimating the spatial autocorrelation matrix of the complete received signal and by subtracting the estimated spatial autocorrelation matrices of the intracell users. The joint detector eliminates the greatest portion of the intracell interference. Therefore, the spatial autocorrelation matrix $\underline{\underline{B}}$ only contains contributions from co-channel interference (CCI) caused by mobile stations (MS) from other cells.

Orthogonal Frequency Division Multiplex (OFDM)

[0047] If a conventional single-carrier transmission system is applied in an environment with said severe transmission conditions, the channel equalization, which is supposed to eliminate the influence of the radio channel as far as possible, can be very extensive. The choice of an appropriate modulation technique for wireless data communication is therefore a critical issue due to the adverse influence of the dispersive and mostly time-variant mobile radio channel. In recent years, the interest in multi-carrier modulation for wireless transmission has been revived, whereas in former times the practicality of this concept appeared to be limited.

[0048] An approach to multi-carrier modulation which can easily be realized is Orthogonal Frequency Division Multiplexing (OFDM). OFDM offers advantages in transmission over severe multipath channels, so that there is an increased interest in applying OFDM in high-rate mobile or portable data transmission today.

[0049] Conventional single-carrier modulation methods for the transmission at high symbol rates experience a severe limitation in time-dispersive and frequency-selective channels due to their sensitivity to intersymbol interference (ISI). To handle ISI, usually the entire bandwidth of the single-carrier signal has to be (adaptively) equalized by quite complex time-domain channel equalizers, like Viterbi equalizers. Thereby, the complexity of a channel equalizer increases with the number of the ISIs which have to be eliminated. If a high data rate R of about 10^7 modulation symbols per second is transmitted over a radio channel having a maximum delay τ_{max} of 10 µs, ISIs can arise extending over 100 modulation symbols. A corresponding equalizer can be too expensive for an implementation. In addition, it is conceivable that the adaptation of the filter coefficients to a time-variant mobile radio channel would show an unstable behavior. Moreover, together with channel coding, reliability information on the equalized channel symbols is desired. Especially, if the channels are as difficult as in mobile communications, channel estimation is complex.

[0050] When operating in a frequency-selective environment, the best interference suppresion performance is generally achieved when wideband algorithms are employed by the adaptive antenna array. However, many wideband adaptive antenna algorithms employ a Fast Fourier Transform (FFT) processing, which is also commonly used in an OFDM demodulator to produce the narrow-band subcarriers of a received OFDM signal. Therefore, a natural way to merge adaptive antennas with OFDM is to employ narrowband adaptive array techniques on the demodulated subcarriers in the OFDM receiver.

[0051] In a pilot-assisted OFDM system, reference-signal-based adaptive array algorithms are appropriate for interference suppression and equalization. Algorithms of this type operate on the pilot symbols transmitted by the desired user to produce a weight vector that attempts to minimize the mean square error between the array output and the known pilot sequence. For best performance, these algorithms need the channel and signal characteristics to be relatively constant over the interval in which said weights are computed and applied to the received signal. Algorithms which can not track channel variations suffer from a significant degradation in the bit error rate (BER) and the signal-to-interference-plus-noise ratio (SINR).

[0052] In an OFDM system employing an adaptive antenna alogorithm on the baseband subcarriers produced by the FFT demodulator, it is difficult to guarantee that the channel will be constant over the intervals in which the weights are computed and applied to the array data. Although each subcarrier can be assumed to be flat faded, the presence of delay spread on the channel can cause significant decorrelation in the fading processes on different subcarriers within a time-frequency slot. Even in a fixed wireless access system, temporal variations in the channel can and will occur due to the motion of any surrounding objects in the system.

Prior patent applications concerning related tasks

[0053] The invention proposed in the European patent application No. 00 118 418 relates to a communication device for receiving and transmitting OFDM signals in a wireless communication system. The system further comprises diversity antenna means including a plurality of antenna elements, means for individually adjusting the amplitude of at least one subcarrier signal of said OFDM signal to be transmitted for each of said antenna elements in accordance with measured attenuation information. Thereby, a higher amplitude is given to each subcarrier of said OFDM transmission signal if said measurement indicates a lower attenuation of the associated transmission channel, and vice versa. In this way, an unnecessary transmission of energy on severely distorted transmission channels can be avoided.

[0054] From the European patent application No. 00 125 436 a method for adjusting the transmission characteristics

of subcarriers of a multi-carrier transmission system using a plurality of antenna elements is known, for example in case of an OFDM multicarrier transmission system. Thereby, the power and the phase of the subcarriers can be adapted. For this purpose, the power of each subcarrier is distributed by means of a weighting unit. Additionally, said subcarriers can be further pre-equalized by dividing the respective subcarrier signal by the sum of the squared magnitude of the frequency channel characteristics of all subcarrier signals or a frequency characteristic of the selected antenna element.

[0055] In the European patent application No. 00 125 435 an adaptive loading calculation and signaling scheme is disclosed for being applied to wireless multi-carrier transmission systems, wherein an Adaptive Loading Calculation block calculates loading tables which contain one entry for each data subcarrier. Thereby, fading channel profile information is used to detect the power of the current fading on each subcarrier. After the subcarriers have been sorted according to their power levels, subcarriers with high power levels use an higher modulation scheme as the originally selected ones, whereas simultaneously subcarriers with low power levels use a lower modulation scheme.

[0056] In the European patent application No. 0 982 875 a method and an apparatus for increasing system capacity and mitigating negative effects caused by high-power users in a mixed-rate CDMA system are disclosed. Thereby, improvements are obtained by applying multi-user detection, antenna array processing or a combination of both techniques to explicitly cancel or attenuate only said high-power users. In this approach a combination of said multi-user detection technique and said antenna array processing technique can be performed to recover system capacity appropriated by the high-power users.

[0057] The US patent application No. 4,353,119 pertains to a receiver system, wherein complex-valued weighting vectors are applied to weight signals from N omnidirectional auxiliary antennas connected to each other. The weighted outputs are then summed with the signal from the main antenna of said receiver system to suppress undesired sidelobe effects caused by the same interferer signal. Moreover, the underlying invention presents a Batch Covariance Relaxation (BCR) approach to solve a complex system of N linear equations in N unknowns involving a Hermitian matrix.

[0058] In the US patent application No. 6,141,393 a method and a device used for a communication system including a receiver having a plurality of adaptive antennas for receiving a plurality of informations bursts transmitted by at least one. transmitting user device is disclosed. Thereby, the information bursts contain a number of data symbols and a pilot symbol sequence of content known at both the transmitting user device and the receiver. Additionally, an error signal between a simulated received pilot signal and the received pilot symbol sequence is calculated, and a channel model sequence is computed, wherein the power of the error signal is minimized and the channel transfer function is computed by weighting predetermined basis functions.

[0059] The US patent application No. 5,694,416 relates to an equipment and methods for increasing the signal-to-interference-plus-noise ratio (SINR) of global positioning system (GPS) receivers. Therein, a navigation satellite receiver for receiving GPS signals being able to suppress interference and enhance satellite signals by using differences in their spatial positions is disclosed. Said receiver comprises four adaptive antennas being arranged in a spatial array. A Code Gated Maximum Likelihood (CGML) technique is applied to whiten predetermined aperture estimates by multiplying these estimates with the mathematical inverse Cholesky factor of the interference data, in order to maximize the ratio of the GPS signal power to the interference power.

DEFICIENCIES AND DISADVANTAGES OF THE KNOWN SOLUTIONS OF THE PRESENT STATE OF THE ART

[0060] As mentioned above, each of the applied interference cancellation techniques is optimized to a specific purpose, and thus it contains certain limitations.

[0061] For example, the European patent application No. 0 982 875 is mainly related to CDMA systems, and the US patent application No. 5,694,416 pertains to GPS systems.

[0062] Using adaptive array antenna schemes or antenna diversity is a conventional method to accommodate large numbers of mobile terminals (MTs) in a cell. Methods of generating antenna weighting vectors according to the present state of the art, which are used to maximize the carrier-to-interference ratio (CIR), normally apply complicated matrix-vector calculations such as the Batch Covariance Relaxation (BCR) approach as described in the US patent application No. 4,353,119, or the Code Gated Maximum Likelihood (CGML) technique as described in the US patent application No. 5,694,416, in which an inverse matrix calculation of the autocorrelation matrix is required.

[0063] Particularly, in multi-carrier systems such as OFDM an antenna weighting vector calculation is performed for each subcarrier. Hence, the total number of necessary calculations is huge and a simpler method is required.

OBJECT OF THE UNDERLYING INVENTION

[0064] In view of the explanations mentioned above it is the object of the invention to propose a simplified low-cost and low-effort interference cancellation technique for a high-speed wireless multi-carrier system. The technique should allow to estimate and equalize the impairments of the received signal caused by the time-variant multipath fading

channel, in order to improve the transmission quality of the system.

[0065] This object is achieved by means of the features of the independent claims. Advantageous features are defined in the dependent claims.

SUMMARY OF THE INVENTION

[0066] The underlying invention describes a low-cost and low-effort solution for an interference cancellation technique applied to a high-speed wireless multi-carrier system for supporting mobile applications within environments being severely impaired by the time-varying multipath fading behavior of the mobile radio channel. Thereby, the vector channel model is applied.

[0067] The N-dimensional signal channel vector of the k-th subcarrier (k = 0, 1, ..., K-1) of a transmitted OFDM signal, given by

$$\underline{a}_k(n) = [a_{k,0}(n), \ldots, a_{k,i}(n), \ldots, a_{k,(N-1)}(n)]^T \in C^N$$

$$\text{for } i = 0,1,\ldots,N-1,$$

wherein n denotes the discrete time variable, contains the complex-valued channel impulse response of all N antenna elements. The N-dimensional complex-valued interference channel vector of the k-th subcarrier is given by

$$\underline{b}_k^{(j)}(n) = [b_{k,0}^{(j)}(n), \ldots, b_{k,i}^{(j)}(n), \ldots, b_{k,(N-1)}^{(j)}(n)]^T \in C^N$$

$$\text{for } i = 0,1,\ldots,N-1 \text{ and } j = 0, 1, \ldots, M-1,$$

wherein j denotes the index of the interferer. Thereby, the signal channel vector $\underline{a}_k(n)$ can be estimated from pilot symbols during the receive mode of Time Division Duplex (TDD). The interference channel vector $\underline{b}_k^{(j)}(n)$ should also be estimated as a required signal. In this case, the signal channel estimation for both required and interference signal should be done separately.

[0068] After combining the required signal from all N antenna elements with the N-dimensional signal channel vector of the k-th subcarrier, given by

$$\underline{w}_k(n) = [w_{k,0}(n), \ldots, w_{k,i}(n), \ldots, w_{k,(N-1)}(n)]^T \in C^N$$

$$\text{for } i = 0, 1, \ldots, N-1,$$

the power of the combined signal - omitting the time variance represented by the discrete time variable n - can be written as

$$E\{|\underline{a}_k^H \underline{w}_k|^2\} = E\{(\underline{a}_k^H \underline{w}_k)^H (\underline{a}_k^H \underline{w}_k)\} = \underline{w}_k^H E\{\underline{a}_k \underline{a}_k^H\} \underline{w}_k = \underline{w}_k^H \underline{\underline{A}}_k \underline{w}_k, \quad (1)$$

wherein

'H' denotes the complex conjugate transpose operation (Hermitian transpose), and

E{·} denotes the expected value of its argument.

[0069] The complex-valued N×N-dimensional desired spatial covariance matrix $\underline{\underline{A}}_k$ is defined as the outer product of the signal channel vectors:

$$\underline{\underline{A}}_k := E\{\underline{a}_k \underline{a}_k^H\} \in C^{N \times N}. \quad (2)$$

[0070] As can be taken from the previous European patent application No. 00 118 418 (the disclosure of which is hereby incorporated by reference), the channel estimate vector $\underline{a}_k$ can also be employed as most adequate antenna weighting vector $\underline{w}_k$ to maximize the required signal power $\underline{w}_k^H \underline{\underline{A}}_k \underline{w}_k$.

[0071] To maximize the carrier-to-interference ratio (CIR), the antenna weighting vector $\underline{w}_k$ is chosen in such a way that the Rayleigh quotient

$$\Gamma_k := \frac{\underline{w}_k^H \underline{\underline{A}}_k \underline{w}_k}{\underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k} \in C, \tag{3}$$

which corresponds to the carrier-to-interference ratio (CIR), can be maximized. Thereby, the denominator denotes the power of the interference which can be calculated as follows:

$$E\left\{\left|\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)^H \underline{w}_k\right|^2\right\} = E\left\{\left[\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)^H \underline{w}_k\right]^*\left[\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)^H \underline{w}_k\right]\right\}$$

$$= E\left\{\left[\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)^H \underline{w}_k\right]^H\left[\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)^H \underline{w}_k\right]\right\} = \underline{w}_k^H E\left\{\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)^H\right\}\underline{w}_k$$

$$= \underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k \quad (4)$$

(using $z^* z \equiv z^H z = \mathrm{Re}\{z\}^2 + \mathrm{Im}\{z\}^2 = |z|^2 \; \forall \; z \in C$),
wherein M is the number of interferers and the complex-valued N×N-dimensional undesired spatial covariance matrix $\underline{\underline{B}}_k$ is defined as the outer product of the summed interference channel vectors:

$$\underline{\underline{B}}_k := \begin{pmatrix} \underline{\underline{B}}_k' & \underline{\underline{0}} \\ \underline{\underline{0}} & \underline{\underline{0}} \end{pmatrix} \in C^{N \times N} \text{ with } \underline{\underline{B}}_k' := E\left\{\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)\left(\sum_{j=0}^{M-1}\underline{b}_k^{(j)}\right)^H\right\} \in C^{M \times M}. \quad (5)$$

Usually, the Rayleigh quotient Γ_k is only defined when its denominator yields a value that is not zero, which implies that the matrix $\underline{\underline{B}}_k$ should be positive definite and, consequently, all eigenvalues of $\underline{\underline{B}}_k$, given by $\lambda_0, \lambda_1, \ldots, \lambda_{(N-1)}$, are greater than zero:

$$\forall \; \underline{w}_k \neq \underline{0}: \; \underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k > 0 \Leftrightarrow \bigwedge_{i=0}^{N-1} (\lambda_i > 0). \quad (6a)$$

[0072] In the underlying invention it is assumed that $\underline{\underline{B}}_k$ is not positive definite (that means $\underline{\underline{B}}_k$ is negative semi-definite):

$$\exists \; \underline{w}_k \neq \underline{0}: \underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k \leq 0 \Leftrightarrow \exists \; \lambda_i \leq 0, \tag{6b}$$

thereby including the cases that the Rayleigh coefficient Γ_k is infinite ($\Gamma_k \to -\infty$). It is further assumed that the matrix $\underline{\underline{B}}_k$ has at least one eigenvector $\lambda_i = 0$ and one dimension of null space. These cases can occur when the matrix $\underline{\underline{B}}_k$ has not full rank (rank $\underline{\underline{B}}_k' = M < N =$ rank $\underline{\underline{B}}_k$) since the number of interferers M is less than the number of antennas N. This means in example: If the number of interferes M is greater than the number of antennas N, only $M^* := N-1$ major interferers can maximally be chosen.

[0073] As mentioned above, one of the objects of this invention is to maximize the carrier-to-interference ratio (CIR) denoted by the Rayleigh quotient Γ_k. For this purpose the denominator $\underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k$ of Γ_k is minimized first. After that, the

antenna weighting vector $\underline{w}_k$ is chosen in such a way that the numerator $\underline{w}_k^H \underline{\underline{A}}_k \underline{w}_k$ of Γ_k becomes a maximum, thereby keeping the denominator $\underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k$ of Γ_k at the already obtained minimum.

[0074] The complex-valued antenna weighting vector $\underline{w}_k^{opt}$ maximizing said Rayleigh quotient Γ_k is given by

$$\underline{w}_k^{opt} = \arg \max_{\underline{w}_k} \{\Gamma_k\} = \arg \max_{\underline{w}_k} \left\{ \frac{\underline{w}_k^H \underline{\underline{A}}_k \underline{w}_k}{\underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k} \right\} \quad (\text{for } \underline{w}_k \neq \underline{0}). \quad (7a)$$

Equation (7a) leads to a generalized eigenvalue problem, given by

$$\underline{\underline{A}}_k \underline{q}_k = \lambda_k \cdot \underline{\underline{B}}_k \underline{q}_k. \quad (7b)$$

[0075] Thereby, it can be shown that the antenna weighting vector $\underline{w}_k^{opt}$ that maximizes the CIR given by said Rayleigh quotient Γ_k in equation (7a) is the so-called dominating generalized eigenvector $\underline{q}_k^{dom}$ associated with the largest generalized eigenvalue λ_k^{max} of the matrix pair $[\underline{\underline{A}}_k, \underline{\underline{B}}_k]$.

[0076] In order to keep the power $\underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k$ of the interference at a minimum, the antenna vector $\underline{w}_k$ is chosen from the orthogonal complement $\beta_k^{\perp}$ of the space

$$\beta_k := \text{span}\{\underline{b}_k^{(j)} \mid 0 \leq j \leq M-1\} \subset C^N, \quad (8)$$

which is spanned by the interference vectors $\underline{b}_k^{(j)}$, wherein $\beta_k^{\perp}$ is defined as follows:

$$\beta_k^{\perp} := \{\underline{w}_k \mid (\underline{w}_k \in C^N) \wedge (\underline{w}_k \perp \underline{b}_k^{(j)} \ \forall \ \underline{b}_k^{(j)} \in \beta_k)\}, \quad (9)$$

wherein $\underline{w}_k \perp \underline{b}_k^{(j)} \Leftrightarrow \underline{w}_k^H \underline{b}_k^{(j)} = 0$.

[0077] Then, in order to maximize the signal power $\underline{w}_k^H \underline{\underline{A}}_k \underline{w}_k$ keeping the power of the interference $\underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k$ at a minimum, the antenna weighting vector $\underline{w}_k$ is defined as the projection of the signal channel vector $\underline{a}_k$ onto the orthogonal complement of the interference channel vector $\underline{b}_k^{(j)}$.

[0078] In case of one interferer (M = 1), antenna weighting vectors $\underline{w}_k$ obtained by means of the following equation maximize the carrier-to-interference ratio (CIR):

$$\text{Step \#0: } \underline{w}_k^{(0)} := \underline{a}_k; \quad (10a)$$

$$\begin{aligned}\text{Step \#1:} \quad \underline{w}_k \equiv \underline{w}_k^{(1)} &:= \underline{w}_k^{(0)} - \text{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(0)}\right) \\ &= \underline{w}_k^{(0)} - \underline{w}_k^{(0)H} \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\|\underline{b}_k^{(0)}\|_2^2}; \quad (10b)\end{aligned}$$

$$\text{with } \text{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(0)}\right) := \underline{w}_k^{(0)H} \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\|\underline{b}_k^{(0)}\|_2^2} \quad (10c).$$

and the Euclidian distance

$$\| \underline{b}_k^{(0)} \|_2 := \sqrt{\underline{b}_k^{(0)H} \underline{b}_k^{(0)}} = \sqrt{\sum_{i=0}^{N-1} b_{k,n}^{(0)\,2}} \,.$$

[0079] Equation (10c) can easily be proved by means of the following geometrical facts (assuming that $\underline{w}_k^{(0)} \neq \underline{0}$ and $\underline{b}_k^{(0)} \neq \underline{0}$) :

$$\sphericalangle\left(\underline{w}_k^{(0)}, \underline{b}_k^{(0)}\right) = \arccos\left(\frac{\underline{w}_k^{(0)H} \underline{b}_k^{(0)}}{\| \underline{w}_k^{(0)} \|_2 \cdot \| \underline{b}_k^{(0)} \|_2}\right) =$$

$$= \arccos\left(\frac{\sum_{i=0}^{N-1} w_{k,i}^{(0)*} \cdot b_{k,i}^{(0)}}{\sqrt{\sum_{i=0}^{N-1} w_{k,i}^{(0)\,2}} \cdot \sqrt{\sum_{i=0}^{N-1} b_{k,i}^{(0)\,2}}}\right), \qquad \text{(11a)}$$

$$\sphericalangle\left(\underline{w}_k^{(0)}, \underline{p}_k^{(0)}\right) = \arccos\left(\frac{\| \underline{p}_k^{(0)} \|_2}{\| \underline{w}_k^{(0)} \|_2}\right) = \arccos\left(\frac{\sqrt{\sum_{i=0}^{N-1} p_{k,i}^{(0)\,2}}}{\sqrt{\sum_{i=0}^{N-1} w_{k,i}^{(0)\,2}}}\right), \qquad \text{(11b)}$$

$$\sphericalangle\left(\underline{w}_k^{(0)}, \underline{b}_k^{(0)}\right) \equiv \sphericalangle\left(\underline{w}_k^{(0)}, \underline{p}_k^{(0)}\right), \qquad \text{(11c)}$$

$$\underline{e}_{b,k}^{(0)} = \frac{\underline{b}_k^{(0)}}{\left\| \underline{b}_k^{(0)} \right\|_2} \quad \text{(with } \|\underline{e}_{b,k}^{(0)}\|_2 = 1\text{)}, \qquad \text{(11d)}$$

and

$$\underline{p}_k^{(0)} = \left\| \underline{p}_k^{(0)} \right\|_2 \cdot \underline{e}_{b,k}^{(0)}, \qquad \text{(11e)}$$

wherein

$\sphericalangle\left(\underline{w}_k^{(0)}, \underline{b}_k^{(0)}\right) \in [0°;180°]$ denotes the angle between $\underline{w}_k^{(0)}$ and $\underline{b}_k^{(0)}$,

$\sphericalangle\left(\underline{w}_k^{(0)}, \underline{p}_k^{(0)}\right) \in [0°;180°]$ denotes the angle between $\underline{w}_k^{(0)}$ and $\underline{p}_k^{(0)}$,

$\underline{e}_k^{(0)}$ is the unit vector in the direction of $\underline{b}_k^{(0)}$,

$$\underline{p}_k^{(0)} \equiv \mathrm{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(0)}\right) \quad \text{is the projection of } \underline{w}_k^{(0)} \text{ onto } \underline{b}_k^{(0)},$$

'*' denotes the complex conjugate operation,
'H' denotes the complex conjugate transpose operation, and
$\|.\|_2$ denotes the Euclidian length of a vector argument.

[0080] After the last step, when the projection of the signal channel vector $\underline{a}_k$ onto the orthogonal complement of all interference channel vectors $\underline{b}_k^{(j)}$ (for $j \in \{0,1\}$) is finished, the obtained antenna weighting vector $\underline{w}_k^{(1)}$ should be normalized again:

$$\underline{w}_k := \frac{\underline{w}_k^{(1)}}{\left\|\underline{w}_k^{(1)}\right\|_2}.$$

[0081] If there are two or more interferers ($M \geq 2$) with known interference channel vectors, then the antenna weighting vector $\underline{w}_k$ can be chosen from the orthogonal complement of all M interference channel vectors, until the number of intereferers M is smaller than the number of antennas N.

[0082] The interference channel vectors $\underline{b}_k^{(j)}$ are assumed to be linearly independent, but normally they are not orthogonal to each other. Therefore, an orthogonalization procedure of the respective interference channel vectors $\underline{b}_k^{(j)}$ should be performed before calculating the antenna weighting vectors $\underline{w}_k$.

[0083] The following equations yield the antenna weighting vectors $\underline{w}_k$ which are needed to achieve a maximization of the carrier-to-interference ratio (CIR) in case of two or more interferers ($M \geq 2$) with known interference channel vectors:

$$\text{Step \#0: } \underline{w}_k^{(0)} := \underline{a}_k; \tag{12a}$$

$$\text{Step \#1:} \quad \underline{w}_k^{(1)} := \underline{w}_k^{(0)} - \mathrm{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(0)}\right) = \underline{w}_k^{(0)} - \underline{w}_k^{(0)H}\underline{b}_k^{(0)} \cdot \underline{u}_k^{(0)}, \tag{12b}$$

wherein

$$\mathrm{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(0)}\right) := \underline{w}_k^{(0)H}\underline{b}_k^{(0)} \cdot \underline{u}_k^{(0)} \quad \text{and} \quad \underline{u}_k^{(0)} := \frac{\underline{b}_k^{(0)}}{\|\underline{b}_k^{(0)}\|_2^2};$$

$$\begin{aligned} \text{Step \#M:} \quad \underline{w}_k &\equiv \underline{w}_k^{(M)} \\ &:= \underline{w}_k^{(0)} - \sum_{j=0}^{M-1} \mathrm{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(j)}\right) = \underline{w}_k^{(0)} - \sum_{j=0}^{M-1} \underline{w}_k^{(0)H}\underline{b}_k^{(j)} \cdot \underline{u}_k^{(j)} \\ &= \underline{w}_k^{(M-1)} - \mathrm{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(M-1)}\right) = \underline{w}_k^{(M-1)} - \underline{w}_k^{(0)H}\underline{b}_k^{(M-1)} \cdot \underline{u}_k^{(M-1)} \end{aligned} \tag{12c}$$

(with M < N), wherein

$$\mathrm{proj}\left(\underline{w}_k^{(0)} \mid \underline{b}_k^{(j)}\right) := \underline{w}_k^{(0)H}\underline{b}_k^{(j)} \cdot \underline{u}_k^{(j)} \quad \text{and} \quad \underline{u}_k^{(j)} := \frac{\underline{b}_k^{(j)}}{\|\underline{b}_k^{(j)}\|_2^2}.$$

[0084] Thereby, $\underline{u}_k^{(j)}$ denotes the normalized orthonormal base of the space $\beta_k^{(j)}$ which is spanned by the interference vectors $\underline{b}_k^{(0)}$, $\underline{b}_k^{(1)}$,..., $\underline{b}_k^{(j)}$, and $\mathrm{proj}(\underline{w}_k^{(0)}|\underline{b}_k^{(j)}$ represents the orthogonal projection of $\underline{w}_k^{(0)}$ onto $\underline{b}_k^{(j)}$.

[0085] After the last step, when the projection of the signal channel vector $\underline{a}_k$ onto the orthogonal complement of all interference channel vectors $\underline{b}_k^{(j)}$ (for $j \in \{0, 1, 2, ..., M\}$) is finished, the obtained antenna weighting vector $\underline{w}_k^{(M)}$ should be normalized again:

$$\underline{w}_k := \frac{\underline{w}_k^{(M)}}{\left\|\underline{w}_k^{(M)}\right\|_2}.$$

[0086] For a generation of the antenna weighting vectors $\underline{w}_k$ for each subcarrier k only selected major interferer(s) can be considered; minor interferer(s) must be neglected as mentioned above. In this case, the normalized orthonormal bases $\underline{u}_k^{(j)}$ can be chosen from previously selected interferer(s).

[0087] Even though the antenna weighting vectors $\underline{w}_k$ are chosen in such a way that the Rayleigh quotient Γ_k becomes infinite, the same ratio including noise power $\sigma_k^2 \cdot I$ for the k-th subcarrier,

$$\Gamma_k' := \frac{\underline{w}_k^H \underline{\underline{A}}_k \underline{w}_k}{\underline{w}_k^H \underline{\underline{B}}_k \underline{w}_k + \sigma_k^2 \cdot I} \in C, \tag{13}$$

which corresponds to the carrier-to-interference-plus-noise ratio (CINR), can be smaller compared with that obtained by means of another antenna weighting vector. This can happen if the signal channel vector $\underline{a}_k$ is highly correlated to one or more interefernce channel vectors $\underline{b}_k^{(j)}$.

[0088] In such cases, the antenna weighting vectors $\underline{w}_k$ for each subcarrier k (for k = 0, 1, ..., K-1) can be found as follows:

[0089] Initially, the antenna weighting vector $\underline{w}_k$ is set to the associated signal channel vector $\underline{a}_k$ (Step #0). Next, $\underline{w}_k$ is gradually modified by removing parts of $\underline{a}_k$ which fall onto each orthogonal base $\underline{u}_k^{(j)}$ of interference (Step #1 to Step #(M-1)). Each time the orthogonal projection

$$\mathrm{proj}\left(\underline{a}_k \,|\, \underline{b}_k^{(j)}\right)$$

of $\underline{a}_k$ onto $\underline{b}_k^{(j)}$ is subtracted as shown above, the Rayleigh quotient Γ_k' is compared before and after said subtraction. If Γ_k' becomes smaller, the subtraction of

$$\mathrm{proj}\left(\underline{a}_k \,|\, \underline{b}_k^{(j)}\right)$$

is abandoned.

[0090] In general, the efficiency of the antenna weighting vectors $\underline{w}_k$ depends on the choice of the normalized orthomormal bases $\underline{u}_k^{(j)}$. In any case, the ratio given by the Rayleigh quotient Γ_k' can be improved, however, since only adequate bases $\underline{u}_k^{(j)}$ are used to generate the antenna weighting vectors $\underline{w}_k$.

[0091] In the following section, an example is given presenting a method how normalized orthonormal bases can be obtained that can be applied to generate the antenna weighting vectors $\underline{w}_k$. In the literature this method is is known as Gram-Schmidt orthogonalization procedure.

[0092] In a first step, the first normalized base $\underline{u}_k^{(0)}$ is derived from the interference channel vector $\underline{b}_k^{(0)}$ of the most dominant interferer:

$$\underline{u}_k^{(0)} := \frac{\underline{b}_k^{(0)}}{\left\|\underline{b}_k^{(0)}\right\|_2^2}. \tag{14a}$$

[0093] Next, the second normalized base $\underline{u}_k^{(1)}$ is chosen from the orthogonal complement of $\underline{u}_k^{(0)}$ and adjusted, in

order to maximize the second dominant interference channel vector $\underline{b}_k^{(1)}$:

$$\text{Step \#1:}\quad \underline{u}_k^{(1)} := \frac{\underline{b}_k^{(1)} - \mathrm{proj}\left(\underline{b}_k^{(1)} \mid \underline{b}_k^{(0)}\right)}{\| \underline{b}_k^{(1)} - \mathrm{proj}\left(\underline{b}_k^{(1)} \mid \underline{b}_k^{(0)}\right) \|_2^2}$$

$$= \frac{\underline{b}_k^{(1)} - \underline{b}_k^{(1)H} \underline{b}_k^{(0)} \cdot \underline{u}_k^{(0)}}{\| \underline{b}_k^{(1)} - \underline{b}_k^{(1)H} \underline{b}_k^{(0)} \cdot \underline{u}_k^{(0)} \|_2^2} \tag{14b}$$

$$\text{wherein } \mathrm{proj}\left(\underline{b}_k^{(1)} \mid \underline{b}_k^{(0)}\right) := \underline{b}_k^{(1)H} \underline{b}_k^{(0)} \cdot \underline{u}_k^{(0)};$$

$$\text{Step \#j:}\quad \underline{u}_k^{(j)} := \frac{\underline{b}_k^{(j)} - \sum_{i=0}^{j-1} \mathrm{proj}\left(\underline{b}_k^{(j)} \mid \underline{b}_k^{(i)}\right)}{\| \underline{b}_k^{(j)} - \sum_{i=0}^{j-1} \mathrm{proj}\left(\underline{b}_k^{(j)} \mid \underline{b}_k^{(i)}\right) \|_2^2}$$

$$= \frac{\underline{b}_k^{(j)} - \sum_{i=0}^{j-1} \underline{b}_k^{(j)H} \underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)}}{\| \underline{b}_k^{(j)} - \sum_{i=0}^{j-1} \underline{b}_k^{(j)H} \underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)} \|_2^2} \tag{14c}$$

$$\text{wherein } \mathrm{proj}\left(\underline{b}_k^{(j)} \mid \underline{b}_k^{(i)}\right) := \underline{b}_k^{(j)H} \underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)};$$

$$\text{Step \#(M-1):}\quad \underline{u}_k^{(M-1)} := \frac{\underline{b}_k^{(M-1)} - \sum_{i=0}^{M-2} \mathrm{proj}\left(\underline{b}_k^{(M-1)} \mid \underline{b}_k^{(i)}\right)}{\| \underline{b}_k^{(M-1)} - \sum_{i=0}^{M-2} \mathrm{proj}\left(\underline{b}_k^{(M-1)} \mid \underline{b}_k^{(i)}\right) \|_2^2}$$

$$= \frac{\underline{b}_k^{(M-1)} - \sum_{i=0}^{M-2} \underline{b}_k^{(M-1)H} \underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)}}{\| \underline{b}_k^{(M-1)} - \sum_{i=0}^{M-2} \underline{b}_k^{(M-1)H} \underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)} \|_2^2} \tag{14d}$$

$$\text{wherein } \mathrm{proj}\left(\underline{b}_k^{(M-1} \mid \underline{b}_k^{(i)}\right) := \underline{b}_k^{(M-1)H} \underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)}.$$

[0094] In order to calculate this set of normalized orthonormal bases, $\underline{U}_k := [\underline{u}_k^{(0)}, \ldots, \underline{u}_k^{(j)}, \ldots, \underline{u}_k^{(M-1)}] \in C^{M \times M}$, in a more efficient way, the algorithm can be simplified as follows:

$$\text{Step \#0:}\quad \underline{u}_k^{(0)} := \frac{\underline{b}_k^{(0)}}{\underline{b}_k^{(0)H} \underline{b}_k^{(0)}}. \tag{15a}$$

$$\text{Step \#1:} \quad \underline{b}_{k,TMP}^{(1)} := \underline{b}_k^{(1)} - \text{proj}\left(\underline{b}_k^{(1)} \mid \underline{b}_k^{(0)}\right) \text{ and}$$

$$\underline{u}_k^{(1)} := \frac{\underline{b}_{k,TMP}^{(1)}}{\underline{b}_{k,TMP}^{(1)\,H}\,\underline{b}_{k,TMP}^{(1)}}, \tag{15b}$$

$$\text{wherein } \text{proj}\left(\underline{b}_k^{(1)} \mid \underline{b}_k^{(0)}\right) := \underline{b}_k^{(1)H}\underline{b}_k^{(0)} \cdot \underline{u}_k^{(0)};$$

$$\text{Step \#j:} \quad \underline{b}_{k,TMP}^{(j)} := \underline{b}_k^{(j)} - \sum_{i=0}^{j-1} \text{proj}\left(\underline{b}_k^{(j)} \mid \underline{b}_k^{(i)}\right) \text{ and}$$

$$\underline{u}_k^{(j)} := \frac{\underline{b}_{k,TMP}^{(j)}}{\underline{b}_{k,TMP}^{(j)\,H}\,\underline{b}_{k,TMP}^{(j)}}, \tag{15c}$$

$$\text{wherein } \text{proj}\left(\underline{b}_k^{(j)} \mid \underline{b}_k^{(i)}\right) := \underline{b}_k^{(j)H}\underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)};$$

$$\text{Step \#(M-1):} \quad \underline{b}_{k,TMP}^{(M-1)} := \underline{b}_k^{(M-1)} - \sum_{i=0}^{M-2} \text{proj}\left(\underline{b}_k^{(M-1)} \mid \underline{b}_k^{(i)}\right)$$

$$\text{and } \underline{u}_k^{(M-1)} := \frac{\underline{b}_{k,TMP}^{(M-1)}}{\underline{b}_{k,TMP}^{(M-1)\,H}\,\underline{b}_{k,TMP}^{(M-1)}}, \tag{15d}$$

$$\text{wherein } \text{proj}\left(\underline{b}_k^{(M-1} \mid \underline{b}_k^{(i)}\right) := \underline{b}_k^{(M-1)H}\underline{b}_k^{(i)} \cdot \underline{u}_k^{(i)}.$$

[0095] This antenna weighting vector generation method is applied to each subcarrier k of the OFDM multi-carrier signal, thereby optimizing the total OFDM multi-carrier signal. As can be seen from equations (15a) to (15d), the underlying invention manages with one inner product operation and one vector subtraction for each interference cancellation - in contrast to the present state of the art.

[0096] Another advantage of the underlying invention is that it the above described method can easily and profitably be applied to Wireless Local Area Networks (W-LANs), in which high bit rates at a rather low mobility of the participants are required.

[0097] The generated antenna weighting vectors $\underline{w}_k$ can also be used for the transmission of TDD signals. By applying the method described above in the transmitter, the access points (APs) can be enabled to extinguish or reduce a signal at specific points where the signal is not required. For example, the signal can be extinguished at specific points where mobile terminals (MTs) are communicating with another AP.

[0098] When antenna weighting vectors as calculated above are applied to an antenna diversity scheme in the transmitter, the distributed transmit signal, $y_k := E\{\underline{w}_k^H \underline{x}_k\}$ $(0 \leq k \leq K-1)$, can also be divided by the inner product of the antenna weighting vector $\underline{w}_k$ with the signal channel vector $\underline{a}_k$:

$$y_k' = \frac{E\{\underline{w}_k^H \underline{x}_k\}}{E\{\underline{w}_k^H \underline{a}_k\}} = \frac{E\left\{\sum_{i=0}^{N-1} w_{k,i}^* \cdot x_{k,i}\right\}}{E\left\{\sum_{i=0}^{N-1} w_{k,i}^* \cdot a_{k,i}\right\}}. \quad (16)$$

[0099] By applying this operation to each subcarrier k, the receiver is enabled to receive equal subcarrier power. Thereby, the obtained signal y_k' can also be distributed.

[0100] Due to said distribution of the transmitted signal, sometimes a very high transmit power is needed for several subcarriers. However, the total transmit power from all antenna elements of a mobile transmitter can also be limited to a predefined threshold. Instead of limiting this transmit power, the underlying modulation scheme can be replaced by a simpler one, or those subcarriers can not be used at all. Therefore, as can be taken from the previous European patent application No. 00 125 435.8, so-called "load swapping" techniques can be employed. These techniques compensate OFDM symbols represented by a reduced bit sequence by changing the modulation scheme of other subcarriers to a more complex scheme which allows to transmit data at a higher data rate. Thereby, the modulation scheme of subcarriers having a power level higher than a predefined threshold is increased, whereas simultaneously the modulation scheme of subcarriers having a power level lower than a predefined threshold is decreased. Usually, the total number of used subcarriers remains unchanged, and the total number of coded bits per OFDM symbol is maintained.

[0101] The antenna weighting vectors can also advantageously be applied to duplex methods used for mobile radio communication. Thereby, access points (APs) can independently communicate with a plurality of mobile terminals (MTs) at the same frequency and time by using orthogonal antenna weighting vectors for each MT. For the purpose of duplex operation, all orthogonal bases referring to the undesired signal channel vector space should be removed from the required signal channel vector $\underline{a}_k$. Theoretically, one AP can accommodate as many MTs as there are antenna elements.

BRIEF DESCRIPTION OF THE CLAIMS

[0102] In general, those skilled in the art will readily recognize that the realization of the underlying invention is not restricted to the above-described examples. Many modifications and variations may be made to the embodiments of the underlying invention disclosed herein without substantially departing from the scope of the invention as defined in the following claims.

[0103] The independent patent claim 1 and the dependent claims 2 to 11 refer to a mobile transmission system comprising at least one mobile transmitter and/or at least one mobile receiver sharing a plurality of mobile radio cells for supporting wireless communication over a mobile radio channel by means of a pilot-assisted wireless multi-carrier system. Said multi-carrier system comprises:

- diversity antennas or adaptive antenna arrays including at least two antenna elements for transmitting and/or receiving modulated radio signals in order to accommodate a large number of mobile terminals within each of said mobile radio cells,

- means for performing an OFDM multi-carrier modulation and/or an OFDM multi-carrier demodulation, respectively, in which each OFDM signal is composed of at least one subcarrier signal each being assigned to a respective transmission channel of said pilot-assisted wireless multi-carrier system, and

- means for individually adjusting the amplitude of at least one subcarrier signal of an OFDM tranmission signal to be transmitted for each of said antenna elements in order to maximize the carrier-to-interference-plus-noise ratio (CINR) for each subcarrier of said mobile transmission system given by the Rayleigh quotients for each subcarrier, which can be calculated by means of the autocorrelation matrix of the signal channel vectors $\underline{a}_k$, the interference channel vectors $\underline{b}_k$, said antenna weighting vectors $\underline{w}_k$ and the noise power $\sigma_k^2 \cdot I$ of the mobile radio channel.

[0104] The mobile radio channel of said pilot-assisted wireless multi-carrier system can be modeled by means of signal channel vectors for each subcarrier and linearly independent interference channel vectors for each subcarrier. In order to perform an interference cancellation, a beamforming algorithm is applied, in which antenna weighting vectors are calculated for each subcarrier in such a way that the denominator of the associated Rayleigh quotient becomes a minimum, thereby simultaneously maximizing the numerator of said Rayleigh quotient.

[0105] In addition, the independent patent claim 12 and the dependent claims 13 to 22 relate to a method for supporting a mobile transmission system as described above.
[0106] The independent patent claim 23 refers to a mobile telecommunications device which supports a mobile transmission system according to anyone of the claims 1 to 11.
[0107] Furthermore, the independent patent claim 24 refers to a mobile telecommunications device which comprises a mobile transmitter and/or a mobile receiver designed for supporting a method according to anyone of the claims 12 to 22.
[0108] Finally, the independent patent claim 25 relates to a computer program, which performs, when executed by a processor of a mobile telecommunication device, a method according to anyone of the claims 12 to 22.

BRIEF DESCRIPTION OF THE DRAWINGS

[0109] Further advantages and suitabilities of the underlying invention result from the subordinate claims as well as from the following description of the preferred embodiment of the invention which is depicted in the following drawings. In this case shows

FIG. 1	a block diagram of the receiver comprising means for performing an interference cancellation in accordance with the preferred embodiment of the underlying invention,
FIG. 2	a block diagram of the transmitter comprising means for performing an interference cancellation in accordance with the preferred embodiment of the underlying invention, wherein a beamforming algorithm is used for maximizing the energy transmitted in direction of the desired mobile terminal MT1 and steering nulls in direction of the interfering mobile terminal MT2,
FIG. 3	a situation in which a first access point AP1 has to independently estimate interferences coming from an interfering mobile terminal MT2 in order to steer nulls in direction of said mobile terminal MT2,
FIG. 4	a block diagram for a mobile transmission system supporting wireless communication over a mobile radio channel by means of a pilot-assisted wireless multi-carrier system considering channel estimation and equalization,
FIG. 5a	a detailed view of the OFDM modulator,
FIG. 5b	a detailed view of the OFDM demodulator
FIG. 6	an example for performing a frequency-domain channel estimation in the receiver by means of preamble symbols,
FIG. 7	a scheme for individually applying antenna weighting vectors to all subcarriers,
FIG. 8	a situation in which one access point AP can transmit different signals to two different mobile stations MT1 and MT2, respectively, at the same frequency and time, and
FIG. 9	a flow chart of an example for choosing the optimum antenna weighting vector according to the preferred embodiment of the underlying invention.

DETAILED DESCRIPTION OF THE UNDERLYING INVENTION

[0110] In the following the functions of the features comprised in the preferred embodiment of the underlying invention as depicted in Figs. 1 to 9 are explained.
[0111] Fig. 1 shows a block diagram of the receiver 100 in accordance with the preferred embodiment of the underlying invention. It comprises means for performing a channel estimation 105a-c and channel equalization 109. For the RF signals received at each of the antennas 102a-c the same operations are performed. In the following, the operations performed for the signal received at antenna 102b shall representatively be described. At first, the RF signal received at antenna 102b is submitted to a module 103b which performs a downconversion to the baseband, an analog-digital conversion and a removal of the guard interval preceding said signal. After that, the OFDM demodulation can be performed by means of a combined module 104b comprising a serial-to-parallel converter (S/P), a digital signal processor performing a Fast Fourier Transform (FFT) and a parallel-to-serial converter (P/S). The obtained data vector

element $x_{k,i}$ is then weighted by means of a receive (RX) antenna weighting vector element $w_{k,i}$ calculated by means of an adaptive processor 106. Thereby, said receive (RX) antenna weighting vector element $w_{k,i}$ is supplied by the channel estimator 105b having the data vector element $x_{k,i}$ as input signal. With the aid of the mixers 107a-c and the summation element 108 an inner product is calculated yielding the scalar signal output $y_k := \underline{w}_k^H \underline{x}_k$ which can be submitted to an equalizer 109. Thereby, the receive (RX) antenna weighting vector $\underline{w}_k$ is adaptively updated by means of the adaptive processor 106.

[0112] Fig. 2 exhibits a block diagram of the transmitter 200 comprising means 207 for performing an interference cancellation in accordance with the preferred embodiment of the underlying invention. Thereby, a beamforming algorithm is used for maximizing the energy transmitted in direction of the desired mobile terminal 201a (MT1) and steering nulls in direction of the interfering mobile terminal 201b (MT2). Again, for the RF signals transmitted from each of the antennas 202a-c the same operations are performed. In the following, the operations performed for the signal transmitted from antenna 202b shall representatively be described. At first, the OFDM symbol stream 205 to be transmitted from each antenna 202a-c is weighted with the aid of transmit (TX) antenna weighting vectors obtained from module 207. After that, the OFDM modulation can be performed by means of a combined module 204b comprising a serial-to-parallel converter (S/P), a digital signal processor performing an Inverse Fast Fourier Transform (IFFT) and a parallel-to-serial converter (P/S). After the OFDM modulation a combined module 203b serves to insert a guard interval, to perform a digital-to-analog conversion and a RF upconversion of the signal to be transmitted from antenna 202b.

[0113] Fig. 3 shows a situation 300 in which a first access point 302a (AP1) has to independently estimate interferences coming from an interfering mobile terminal 301b (MT2) in order to steer nulls in direction of said mobile terminal 301b (MT2). Likewise, Fig. 3 shows a situation in which a second access point 302b (AP2) has to independently estimate interferences coming from an interfering mobile terminal 301a (MT1) in order to steer nulls in direction of said mobile terminal 301a (MT1). Thereby, the access points 302a (AP1) and 302b (AP2) can co-operate by exchanging specific signal bursts comtained in their signal sequences 303a and 303b, respectively.

[0114] Fig. 4 refers to a block diagram 400 for the employed mobile transmission system comprising one mobile transmitter 401 and one mobile receiver 402 for supporting wireless communication over a mobile radio channel by means of a pilot-assisted wireless multi-carrier system considering channel estimation and equalization.

[0115] The input is a binary data stream 404 using any suitable modulation technique. At first, a scrambler 405 is used for randomizing the transmitted input data bits 404 in order to minimize interferences. After being submitted to an encoder 406 and a bit-to-symbol mapper 407 followed by an interleaver 408, the data is then transformed into a multilevel signal to be prepared for an OFDM modulation 409. When the serial data stream is converted to parallel by means of the serial-parallel converter 409a, the data rate gets reduced by K, where K is the number of parallel subchannels used for the applied OFDM modulation. Hence, these parallel channels are essentially low data rate channels, and since they are narrow-band channels they experience flat fading. This is the greatest advantage of the applied OFDM technique. To obtain orthogonality between the subcarriers, the data symbols are mapped to the subcarriers using an Inverse Fast Fourier Transform (IFFT) performed by a digital signal processor 409b, and reconverted to serial by means of a parallel-serial converter 409c. After OFDM modulation a guard interval is inserted with the aid of the guard interval insertion unit 410. The use of a cyclic prefix (CP) instead of a plain guard interval simplifies the channel equalization in the receiver. Moreover, it is advantageous to maintain carrier synchronization in the receiver. Next, a signal windowing is performed in order to reduce the out-of-band radiation. Therefore, a raised-cosine window function is normally employed. The signal is then submitted to a digital-analog converter 411 to produce the analog baseband signal, modulated to the RF carrier wave by means of an RF upconversion unit 412a, amplified, and transmitted over the mobile radio channel 403. Thereby, the mobile radio channel 403 is assumed to be an Additive White Gaussian Noise (AWGN) channel. In the receiver the complementary operations are applied in reverse order. At first, the received RF signal is downconverted to the baseband by means of the RF downconverter 414b, and submitted to an analog-digital converter 415. When the guard interval is removed with the aid of the guard interval removal unit 416, the OFDM demodulation 417 can be performed. The data is then transformed into a multilevel signal to be prepared for an OFDM demodulation. After the serial data stream is converted to parallel by means of the serial-parallel converter 417a, all subcarriers are separated by applying a Fast Fourier Transform (FFT) performed by a digital signal processor 417b, and reconverted to serial by means of a parallel-serial converter 417c. After the performance of the channel estimation 418 and equalization 419, the data stream is submitted to a deinterleaver 420 followed by a symbol-to-bit mapper 421, a decoder 422, and a descrambler 423 to obtain the output data bits 424.

[0116] Figs. 5a and 5b exhibit a detailed view of the OFDM modulation 501 performed in the transmitter with multiplexing data symbols and pilot symbols, and the OFDM demodulation 502 performed in the receiver, respectively. Thereby, the data is modulated on $N_{ST} = 64$ subcarriers by means of an Inverse Fast Fourier Transform (IFFT) in the transmitter, and demodulated by means of a Fast Fourier Transform (FFT) in the receiver. The applied OFDM parameters can be taken from the following table:

OFDM Transmission Parameter	Prescribed Value
sampling period Δt	$\Delta t = 50$ ns
sampling rate f_s	$f_s = 1 / \Delta t = 20$ MHz
symbol duration T_s	$T_s = 64 \cdot \Delta t = 3.2\ \mu s$
guard interval T_G (required | optional)	$T_G = 16 \cdot \Delta t = 0.8\ \mu s$ | $T_G = 8 \cdot \Delta t = 0.4\ \mu s$
symbol interval T_{MC} (required | optional)	$T_{MC} = T_S + T_G = 80 \cdot \Delta t = 4.0\ \mu s$ | $T_{MC} = T_S + T_G = 72 \cdot \Delta t = 3.6\ \mu s$
number N_{SD} of subcarriers for data signals	$N_{SD} = 48$
number N_{SP} of subcarriers for pilot signals	$N_{SP} = 4$
total number N_{ST} of subcarriers	$N_{ST} \equiv K = N_{SD} + N_{SP} = 52$
subchannel spacing Δf between the individual subcarriers	$\Delta f = 1 / T_S = 0.3125$ MHz
subchannel spacing B between the outer subcarriers	$B = N_{ST} \cdot \Delta f = 16.25$ MHz
system bandwidth B_{sys}	$B_{sys} = 20$ MHz

[0117] Fig. 6 shows an example 600 for performing a frequency-domain channel estimation in the receiver 100 by means of preamble symbols. Thereby, an alignment of the subcarrier phase is performed with a predetermined pattern, given by the respective receive (RX) antenna weighting vector element $w_{k,i}$ calculated by means of the adaptive processor 106. 601a is an example for an OFDM symbol stream before said alignment of the subcarrier phases; 601b shows the OFDM symbol stream after said alignment.

[0118] Fig. 7 presents a scheme 700 for individually applying antenna weighting vectors to all subcarriers. Thereby, pre-equalizing and load swapping techniques can optionally be added. 701 shows a matrix scheme comprising transmit (TX) antenna weighting vector elements for each subcarrier k (for $0 \le k \le K-1$) and transmit (TX) antenna i (for $0 \le i \le N-1$) to maximize the carrier-to-interference ratio (CIR) given by the Rayleigh quotient Γ_k, or the carrier-to-interference-plus-noise ratio (CINR) given by the Rayleigh quotient Γ_k', respectively. These transmit (TX) antenna weighting vector elements are needed to generate the non-normalized scalar output signals y_k (for $0 \le k \le K-1$) depicted in the vector scheme 702. Alternatively, vector scheme 703 comprising the distributed normalized output power for each subcarrier k to be transmitted in case of applying a magnitude adjustment for limiting said output power (= Option #1) can be employed, or vector scheme 704 comprising the distributed normalized output power for each subcarrier k to be transmitted using a combined threshold and load swapping technique in order to limit said output power (= Option #2).

[0119] Fig. 8 exhibits a situation 800 in which one access point 802 (AP) can transmit different signals to two different mobile stations 801a (MT1) and 801b (MT2), respectively, at the same frequency and time. Therein, two mutually orthogonal antenna weighting vectors are employed to form null and beam, respectively.

[0120] Fig. 9 shows a a flow chart 900 of an example for choosing the optimum antenna weighting vector according to the preferred embodiment of the underlying invention. It contains the algorithm as described above.

[0121] At first, an initialization 901 of the antenna weighting vector is performed by setting the antenna weighting vector to the associated signal channel vector:

$$\underline{w}_k^{opt} := \underline{a}_k.$$

[0122] Next, a calculation 902 of the Rayleigh quotient

$$\Gamma_{k,before}' := \frac{\underline{w}_k^{opt^H} \underline{\underline{A}}_k \underline{w}_k^{opt}}{\underline{w}_k^{opt^H} \underline{\underline{B}}_k \underline{w}_k^{opt} + \sigma_k^2 \cdot I}$$

representing the carrier-to-interference-plus-noise ratio (CINR) before subtracting the projection of the associated signal channel vector $\underline{a}_k$ onto the orthogonal base $\underline{u}_k^{(j)}$ of the associated interference channel vector $\underline{b}_k^{(j)}$ is done. After a reinitialization 903 of the interferer index ($j := 0$), a calculation 904 of a new antenna weighting vector by means of a

first Gram-Schmidt orthogonalization procedure is executed:

$$\underline{w}_k^{try} := \underline{a}_k - \mathrm{proj}\left(\underline{a}_k \mid \underline{b}_k^{(0)}\right) = \underline{a}_k - \underline{a}_k^H \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\underline{b}_k^{(0)H} \underline{b}_k^{(0)}}$$

$$\text{with } \mathrm{proj}\left(\underline{a}_k \mid \underline{b}_k^{(0)}\right) := \underline{a}_k^H \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\underline{b}_k^{(0)H} \underline{b}_k^{(0)}} = \underline{a}_k^H \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\| \underline{b}_k^{(0)} \|_2^2} .$$

[0123] In the next step 905, the Rayleigh quotient

$$\Gamma_{k,after}' := \frac{\underline{w}_k^{try^H} \underline{\underline{A}}_k \underline{w}_k^{try}}{\underline{w}_k^{try^H} \underline{\underline{B}}_k \underline{w}_k^{try} + \sigma_k^2 \cdot I}$$

representing the carrier-to-interference-plus-noise ratio (CINR) after subtracting the projection of the associated signal channel vector $\underline{a}_k$ onto the orthogonal base $\underline{u}_k^{(j)}$ of the associated interference channel vector $\underline{b}_k^{(j)}$ is calculated. Then, in step 906, the Rayleigh quotient before performing step 904 is compared with the Rayleigh quotient after having performed step 904. If the query

$$\text{"}\Gamma_{k,after}' > \Gamma_{k,before}' \text{ ?"}$$

yields 'true', the algorithm continues with step 907. Otherwise, if said query yields 'false', the termination of said algorithm is initiated. Next, in step 907, it has to be checked whether the current interferer index is the last interferer index or not. If the query

$$\text{"}j = N-1 \text{ ?"}$$

yields 'true', an updating 910 of the antenna weighting vector is performed:

$$\underline{w}_k^{opt} := \underline{w}_k^{try},$$

and the termination of said algorithm is initiated. Otherwise, if said query yields 'false', an updating of the Rayleigh quotient calculated before executing step 904 is performed in step 908:

$$\Gamma_{k,before}' := \Gamma_{k,after}',$$

and in a combined step 909 an incrementation of the interferer index ($j := j+1$), and an updating of the antenna weighting vector is performed as shown in step 910. After that, the algorithm starts again with step 904.

[0124] The significance of the symbols designated with reference signs in the figures 1 to 9 can be taken from the appended table of reference signs.

[0125] Table of the depicted features and their corresponding reference signs

No.	Feature
100	block diagram of a mobile receiver comprising means for performing an interference cancellation in accordance with the preferred embodiment of the underlying invention
101a	transmit (TX) antenna of a first mobile terminal (MT1) transmitting desired signals
101b	transmit (TX) antenna of a second mobile terminal (MT2) transmitting interference signals
102a	1st receive (RX) antenna of the mobile receiver 100 receiving signals coming from MT1 and MT2
102b	2nd receive (RX) antenna of the mobile receiver 100 receiving signals coming from MT1 and MT2
102c	N-th receive (RX) antenna of the mobile receiver 100 receiving signals coming from MT1 and MT2
103a	1st combined module of the mobile receiver 100 comprising means for a downconversion, an analog-to-digital conversion (A/D) of the received signal and a guard interval reduction (GIR)
103b	2nd combined module of the mobile receiver 100 comprising means for a downconversion, an analog-to-digital conversion (A/D) of the received signal and a guard interval reduction (GIR)
103c	N-th combined module of the mobile receiver 100 comprising means for a downconversion, an analog-to-digital conversion (A/D) of the received signal and a guard interval reduction (GIR)
104a	1st combined module of the mobile receiver 100 comprising a serial-to-parallel converter (S/P), a digital signal processor performing a Fast Fourier Transform (FFT) and a parallel-to-serial converter (P/S)
104b	2nd combined module of the mobile receiver 100 comprising a serial-to-parallel converter (S/P), a digital signal processor performing a Fast Fourier Transform (FFT) and a parallel-to-serial converter (P/S)
104c	N-th combined module of the mobile receiver 100 comprising a serial-to-parallel converter (S/P), a digital signal processor performing a Fast Fourier Transform (FFT) and a parallel-to-serial converter (P/S)
105a	1st frequency-domain channel estimator of the mobile receiver 100
105b	2nd frequency-domain channel estimator of the mobile receiver 100
105c	N-th frequency-domain channel estimator of the mobile receiver 100
106	adaptive processor of the mobile receiver 100 for calculating receive (RX) antenna weighting vector elements for

No.	Feature
	each subcarrier k (for $0 \le k \le K-1$) and each receive (RX) antenna i (for $0 \le i \le N-1$) to maximize the carrier-to-interference ratio (CIR) given by the Rayleigh quotient Γ_k, or the carrier-to-interference-plus-noise ratio (CINR) given by the Rayleigh quotient Γ_k', respectively
107a	1st mixer of the mobile receiver 100
107b	2nd mixer of the mobile receiver 100
107c	N-th mixer of the mobile receiver 100
108	summation element of the mobile receiver 100
109	combined module of the mobile receiver 100 comprising means for performing an equalization and a demodulation of the received signal
200	block diagram of a mobile transmitter or a transmitting module in an access point (AP) comprising means for performing an interference cancellation in accordance with the preferred embodiment of the underlying invention, wherein a beamforming algorithm is used for maximizing the energy transmitted in direction of the desired mobile terminal MT1 and steering nulls in direction of the interfering mobile terminal MT2
201a	mobile terminal MT1 transmitting desired signals
201b	mobile terminal MT2 transmitting interference signals
202a	1st transmit (TX) antenna of the mobile transmitter 200 transmitting signals to MT1 and MT2
202b	2nd transmit (TX) antenna of the mobile transmitter 200 transmitting signals to MT1 and MT2
202c	N-th transmit (TX) antenna of the mobile transmitter 200 transmitting signals to MT1 and MT2
203a	1st combined module of the mobile transmitter 200 comprising means for a guard interval insertion (GII), a digital-to-analog conversion (D/A) and an upconversion and of the received signal
203b	2nd combined module of the mobile transmitter 200 comprising means for a guard interval insertion (GII), a digital-to-analog conversion (D/A) and an upconversion and of the received signal
203c	N-th combined module of the mobile transmitter 200 comprising means for a guard interval insertion (GII), a digital-to-analog conversion (D/A) and an upconversion and of the received signal
204a	1st combined module of the mobile transmitter 200 comprising a serial-to-parallel converter (S/P), a digital signal processor performing an Inverse Fast Fourier Transform (IFFT) and a parallel-to-serial converter (P/S)
204b	2nd combined module of the mobile transmitter 200 comprising a serial-to-parallel converter (S/P), a digital signal processor performing an Inverse Fast Fourier Transform (IFFT) and a parallel-to-serial converter (P/S)
204c	N-th combined module of the mobile transmitter 200 com-

No.	Feature
	prising a serial-to-parallel converter (S/P), a digital signal processor performing an Inverse Fast Fourier Transform (IFFT) and a parallel-to-serial converter (P/S)
205	module of the mobile transmitter 200 supplying a frequency-domain OFDM symbol stream to be transmitted
206a	1st mixer of the mobile transmitter 200
206b	2nd mixer of the mobile transmitter 200
206c	N-th mixer of the mobile transmitter 200
207	module for calculating transmit (TX) antenna weighting vector elements for each subcarrier k (for $0 \le k \le K-1$) and each transmit (TX) antenna i (for $0 \le i \le N-1$) to maximize the carrier-to-interference ratio (CIR) given by the Rayleigh quotient Γ_k, or the carrier-to-interference-plus-noise ratio (CINR) given by the Rayleigh quotient Γ_k', respectively
207a	vector scheme comprising the calculated (TX) antenna weighting vector elements of transmit (TX) antenna #0 (for $0 \le i \le N-1$) for each subcarrier k (for $0 \le k \le K-1$)
207b	vector scheme comprising the calculated (TX) antenna weighting vector elements of transmit (TX) antenna #1 (for $0 \le i \le N-1$) for each subcarrier k (for $0 \le k \le K-1$)
207c	vector scheme comprising the calculated (TX) antenna weighting vector elements of transmit (TX) antenna #(N-1) (for $0 \le i \le N-1$) for each subcarrier k (for $0 \le k \le K-1$)
300	situation in which a first access point AP1 has to independently estimate interferences coming from an interfering mobile terminal MT2 in order to steer nulls in direction of said mobile terminal MT2
301a	transmit and/or receive (TX/RX) antenna of a first mobile terminal (MT1) transmitting and/or receiving desired signals and/or interference signals
301b	transmit and/or receive (TX/RX) antenna of a second mobile terminal (MT2) transmitting and/or receiving desired signals and/or interference signals
302a	transmit and/or receive (TX/RX) antenna array of a first access point AP1
302b	transmit and/or receive (TX/RX) antenna array of a second access point AP2
303a	received sequence of desired signals (black) and interference signals (hatched) in access point AP1
303b	received sequence of desired signals (black) and interference signals (hatched) in access point AP2
400	block diagram for a mobile transmission system supporting wireless communication over a mobile radio channel by means of a pilot-assisted wireless multi-carrier system considering channel estimation and equalization
401	transmitter of said mobile transmission system 400
402	receiver of said mobile transmission system 400

No.	Feature
403	Additive White Gaussian Noise (AWGN) channel characterized by a severe frequency-selective fading and a time-variant behavior
404	input data bits
405	scrambler used for randomizing the deterministic behavior of predefined pilots in order to minimize interferences
406	encoder (e.g. a convolutional encoder)
407	interleaver
408	bit-to-symbol mapper
409	OFDM modulator
409a	serial-parallel (S/P) converter
409b	digital signal processor performing an Inverse Fast Fourier Transform (IFFT)
409c	parallel-serial (P/S) converter
410	guard interval insertion (GII) and windowing in order to reduce the out-of-band radiation
411	digital-to-analog (D/A) converter
412a	RF upconversion unit
412b	transmitting antenna
413	additive noise
414a	receiving antenna
414b	RF downconversion unit
415	analog-to-digital (A/D) converter
416	guard interval removal (GIR) and windowing in order to reduce the out-of-band radiation
417	OFDM demodulator
417a	serial-parallel (S/P) converter
417b	digital signal processor performing a Fast Fourier Transform (FFT)
417c	parallel-serial (P/S) converter
418	channel estimator
419	channel equalizer
420	symbol-to-bit mapper
421	deinterleaver
422	decoder (e.g. a convolutional decoder)
423	descrambler
424	output data bits
501	detailed view of the OFDM modulator 409
502	detailed view of the OFDM demodulator 417
600	example for performing a frequency-domain channel estimation in the receiver by means of preamble symbols
601a	example for an OFDM symbol stream before the alignment of the subcarrier phases by means of predetermined pilot patterns
601b	example for an OFDM symbol stream after the alignment of the subcarrier phases by means of predetermined pilot patterns
602	detailed view of the receiver 100 showing the frequency-

No.	Feature
	domain channel estimation, which is performed with the aid of preamble symbols contained in the received pilot patterns
602a	combined module of the mobile receiver 100 comprising a serial-to-parallel converter (S/P), a digital signal processor performing a Fast Fourier Transform (FFT) and a parallel-to-serial converter (P/S)
602b	mixer of the mobile receiver 100 performing a phase alignment with a predetermined pilot pattern
700	scheme for individually applying antenna weighting vectors to all subcarriers
701	matrix scheme comprising transmit (TX) antenna weighting vector elements for each subcarrier k (for $0 \leq k \leq K-1$) and each transmit (TX) antenna i (for $0 \leq i \leq N-1$) to maximize the carrier-to-interference ratio (CIR) given by the Rayleigh quotient Γ_k, or the carrier-to-interference-plus-noise ratio (CINR) given by the Rayleigh quotient Γ_k', respectively
702	vector scheme comprising the distributed non-normalized output signal for each subcarrier k (for $0 \leq k \leq K-1$) to be transmitted
703	vector scheme comprising the distributed normalized output power for each subcarrier k (for $0 \leq k \leq K-1$) to be transmitted in case of applying a magnitude adjustment for limiting said output power (= Option #1)
704	vector scheme comprising the distributed normalized output power for each subcarrier k (for $0 \leq k \leq K-1$) to be transmitted in case of applying a combined threshold and load swapping technique in order to limit said output power (= Option #2)
800	situation in which one access point AP can transmit different signals to two different mobile stations MT1 and MT2, respectively, at the same frequency and time
801a	transmit and/or receive (TX/RX) antenna of a first mobile terminal (MT1) transmitting and/or receiving desired signals and/or interference signals
801b	transmit and/or receive (TX/RX) antenna of a second mobile terminal (MT2) transmitting and/or receiving desired signals and/or interference signals
802	transmit and/or receive (TX/RX) antenna array of an access point AP
803	transmitted sequence of desired signals (black) from the access point AP to MT1 and/or MT2
900	flow chart of an example for choosing the optimum antenna weighting vector according to the preferred embodiment of the underlying invention
901	initialization of the antenna weighting vector by setting the antenna weighting vector to the associated signal

No.	Feature
	channel vector: $\underline{w}_k^{opt} := \underline{a}_k$
902	calculation of the Rayleigh quotient $$\Gamma_{k,before}' := \frac{\underline{w}_k^{opt^H} \underline{\underline{A}}_k \underline{w}_k^{opt}}{\underline{w}_k^{opt^H} \underline{\underline{B}}_k \underline{w}_k^{opt} + \sigma_k^2 \cdot I}$$ representing the carrier-to-interference-plus-noise ratio before subtracting the projection of the associated signal channel vector $\underline{a}_k$ onto the orthogonal base $\underline{u}_k^{(j)}$ of the associated interference channel vector $\underline{b}_k^{(j)}$
903	reinitialization of the interferer index: $j := 0$
904	calculation of a new antenna weighting vector by means of a first Gram-Schmidt orthogonalization procedure: $$\underline{w}_k^{try} := \underline{a}_k - \text{proj}\left(\underline{a}_k \mid \underline{b}_k^{(0)}\right) = \underline{a}_k - \underline{a}_k^H \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\underline{b}_k^{(0)^H} \underline{b}_k^{(0)}}$$ with $\text{proj}\left(\underline{a}_k \mid \underline{b}_k^{(0)}\right) := \underline{a}_k^H \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\underline{b}_k^{(0)^H} \underline{b}_k^{(0)}} = \underline{a}_k^H \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\lVert \underline{b}_k^{(0)} \rVert_2^2}$
905	calculation of the Rayleigh quotient $$\Gamma_{k,after}' := \frac{\underline{w}_k^{try^H} \underline{\underline{A}}_k \underline{w}_k^{try}}{\underline{w}_k^{try^H} \underline{\underline{B}}_k \underline{w}_k^{try} + \sigma_k^2 \cdot I}$$ representing the carrier-to-interference-plus-noise ratio after subtracting the projection of the associated signal channel vector $\underline{a}_k$ onto the orthogonal base $\underline{u}_k^{(j)}$ of the associated interference channel vector $\underline{b}_k^{(j)}$
906	query whether the Rayleigh quotient has been increased or not by performing the algorithm above: $\Gamma_{k,after}' > \Gamma_{k,before}'$?
907	query whether the current interferer index is the last interferer index or not: $j = N-1$?
908	updating of the Rayleigh quotient: $\Gamma_{k,before}' := \Gamma_{k,after}'$
909	incrementation of the interferer index: $j := j+1$, and updating of the antenna weighting vector: $\underline{w}_k^{opt} := \underline{w}_k^{try}$
910	updating of the antenna weighting vector: $\underline{w}_k^{opt} := \underline{w}_k^{try}$

Claims

1. A mobile transmission system comprising at least one mobile transmitter (200) and/or at least one mobile receiver (100) sharing a plurality of mobile radio cells for supporting wireless communication over a mobile radio channel by means of a pilot-assisted wireless multi-carrier system, comprising:

- diversity antennas or adaptive antenna arrays (102a-c, 202a-c) including at least two antenna elements for transmitting and/or receiving modulated radio signals,
- means for performing an OFDM multi-carrier modulation (204a-c) and/or demodulation (104a-c), in which each OFDM signal is composed of at least one subcarrier signal each being assigned to a respective transmission channel of said pilot-assisted wireless multi-carrier system,
- means (106, 107a-c, 108) for individually adjusting the amplitude of at least one subcarrier signal of an OFDM tranmission signal to be transmitted for each of said antenna elements (102a-c) in order to maximize the carrier-to-interference-plus-noise ratio (CINR) for each subcarrier of said mobile transmission system given by the Rayleigh quotients for each subcarrier,

characterized by
means for carrying out a beamforming algorithm, in which antenna weighting vectors are calculated for each subcarrier in such a way that the denominator of the associated Rayleigh quotient becomes a minimum.

2. A mobile transmission system according to claim 1,
wherein the antenna weighting vectors for each subcarrier are chosen from the orthogonal complement of the space which is spanned by the associated interference channel vectors.

3. A mobile transmission system according to anyone of the claims 1 or 2,
characterized by
means for performing a first Gram-Schmidt orthogonalization for the interference channel vectors of each subcarrier before calculating the associated antenna weighting vectors for each subcarrier.

4. A mobile transmission system according to claim 3,
wherein the means for performing said first Gram-Schmidt orthogonalization procedure perform the following steps for each subcarrier:

- initialization of the antenna weighting vector by setting the antenna weighting vector to the associated signal channel vector,
- gradual modification of the antenna weighting vector by subtracting the projection of the associated signal channel vector onto the orthogonal base of the associated interference channel vector, in order to obtain an orthogonal set of linearly independent antenna weighting vectors,
- comparation of the the Rayleigh quotient before and after said subtraction,
- abandonment of the the subtraction of said projection if the Rayleigh quotient calculated after having performed said subtraction is smaller than before.

5. A mobile transmission system according to anyone of the claims 3 or 4,
comprising means for the execution of said first Gram-Schmidt orthogonalization procedure for calculating antenna weighting vectors,
wherein a normalized orthogonal base is calculated with the aid of a second Gram-Schmidt orthogonalization procedure in order to improve the efficiency of said first Gram-Schmidt orthogonalization procedure.

6. A mobile transmission system according to anyone of the claims 1 to 5,
comprising means for normalizing the output power for all subcarriers of the mobile transmitter by dividing it by the inner product of the antenna weighting vector and the signal channel vector in order to obtain an equal subcarrier power in the mobile receiver.

7. A mobile transmission system according to claim 6,
wherein the transmitted signal is distributed to all antenna elements of the mobile transmitter after said normalization of the output power for all subcarriers of said mobile transmitter.

8. A mobile transmission system according to anyone of the claims 1 to 7,
wherein the total transmit power from all antenna elements of the mobile transmitter is limited to a predefined upper threshold level.

9. A mobile transmission system according to anyone of the claims 1 to 7,
wherein a load swapping technique is employed in order to compensate OFDM symbols represented by a reduced bit sequence by changing the modulation scheme of other subcarriers to a more complex scheme which allows

to transmit data at a higher data rate.

10. A mobile transmission system according to claim 9, comprising means for performing a load swapping, wherein the modulation scheme of subcarriers having a power level higher than a predefined threshold is increased, whereas the modulation scheme of subcarriers having a power level lower than a predefined threshold is decreased, thereby departing from the applied OFDM modulation scheme.

11. A mobile transmission system according to anyone of the claims 1 to 10, comprising at least one access point (AP) and at least two mobile terminals (MT), wherein orthogonalized antenna weighting vectors are applied to provide a duplex operation in said mobile transmission system in order to enable said access points (AP) to independently communicate with two or more of said mobile terminals (MT) at the same frequency and/or time.

12. A method for operating a mobile transmission system comprising at least one mobile transmitter (200) and/or at least one mobile receiver (100) sharing a plurality of mobile radio cells for supporting wireless communication over a mobile radio channel by means of a pilot-assisted wireless multi-carrier system, comprising:

- diversity antennas or adaptive antenna arrays (102a-c, 202a-c) including at least two antenna elements for transmitting and/or receiving modulated radio signals,
- means for performing an OFDM multi-carrier modulation (204a-c) and/or demodulation (104a-c), in which each OFDM signal is composed of at least one subcarrier signal each being assigned to a respective transmission channel of said pilot-assisted wireless multi-carrier system,
- means (106, 107a-c, 108) for individually adjusting the amplitude of at least one subcarrier signal of an OFDM tranmission signal to be transmitted for each of said antenna elements (102a-c) in order to maximize the carrier-to-interference-plus-noise ratio (CINR) for each subcarrier of said mobile transmission system given by the Rayleigh quotients for each subcarrier,

characterized in that
antenna weighting vectors are calculated for each subcarrier in such a way that the denominator of the associated Rayleigh quotient becomes a minimum, thereby maximizing the numerator of said Rayleigh quotient.

13. A method according to claim 12, wherein the antenna weighting vectors for each subcarrier are chosen from the orthogonal complement of the space which is spanned by the associated interference channel vectors.

14. A method according to anyone of the claims 12 or 13, wherein a first Gram-Schmidt orthogonalization is performed for the interference channel vectors of each subcarrier before calculating the associated antenna weighting vectors for each subcarrier.

15. A method according to claim 14, wherein for the execution of said first Gram-Schmidt orthogonalization procedure for calculating antenna weighting vectors the following steps are performed:

- initialization of the antenna weighting vector by setting the antenna weighting vector to the associated signal channel vector,
- gradual modification of the antenna weighting vector by subtracting the projection of the associated signal channel vector onto the orthogonal base of the associated interference channel vectorin order to obtain an orthogonal set of linearly independent antenna weighting vectors,
- comparation of the the Rayleigh quotient before and after said subtraction,
- abandonment of the the subtraction of said projection if the Rayleigh quotient calculated after having performed said subtraction is smaller than before.

16. A method according to anyone of the claims 14 or 15, wherein a normalized orthogonal base is calculated with the aid of a second Gram-Schmidt orthogonalization procedure in order to improve the efficiency of said first Gram-Schmidt orthogonalization procedure.

17. A method according to anyone of the claims 12 to 16, wherein the output power for all subcarriers of the mobile transmitter is normalized by dividing it by the inner product

of the antenna weighting vector and the signal channel vector in order to obtain an equal subcarrier power in the mobile receiver.

18. A method according to claim 17, wherein the transmitted signal is distributed to all antenna elements of the mobile transmitter after said normalization of the output power for all subcarriers of said mobile transmitter.

19. A method according to anyone of the claims 12 to 18, wherein the total transmit power from all antenna elements of the mobile transmitter is limited to a predefined upper threshold level.

20. A method according to anyone of the claims 12 to 18, wherein a load swapping technique is employed in order to compensate OFDM symbols represented by a reduced bit sequence by changing the modulation scheme of other subcarriers to a more complex scheme which allows to transmit data at a higher data rate.

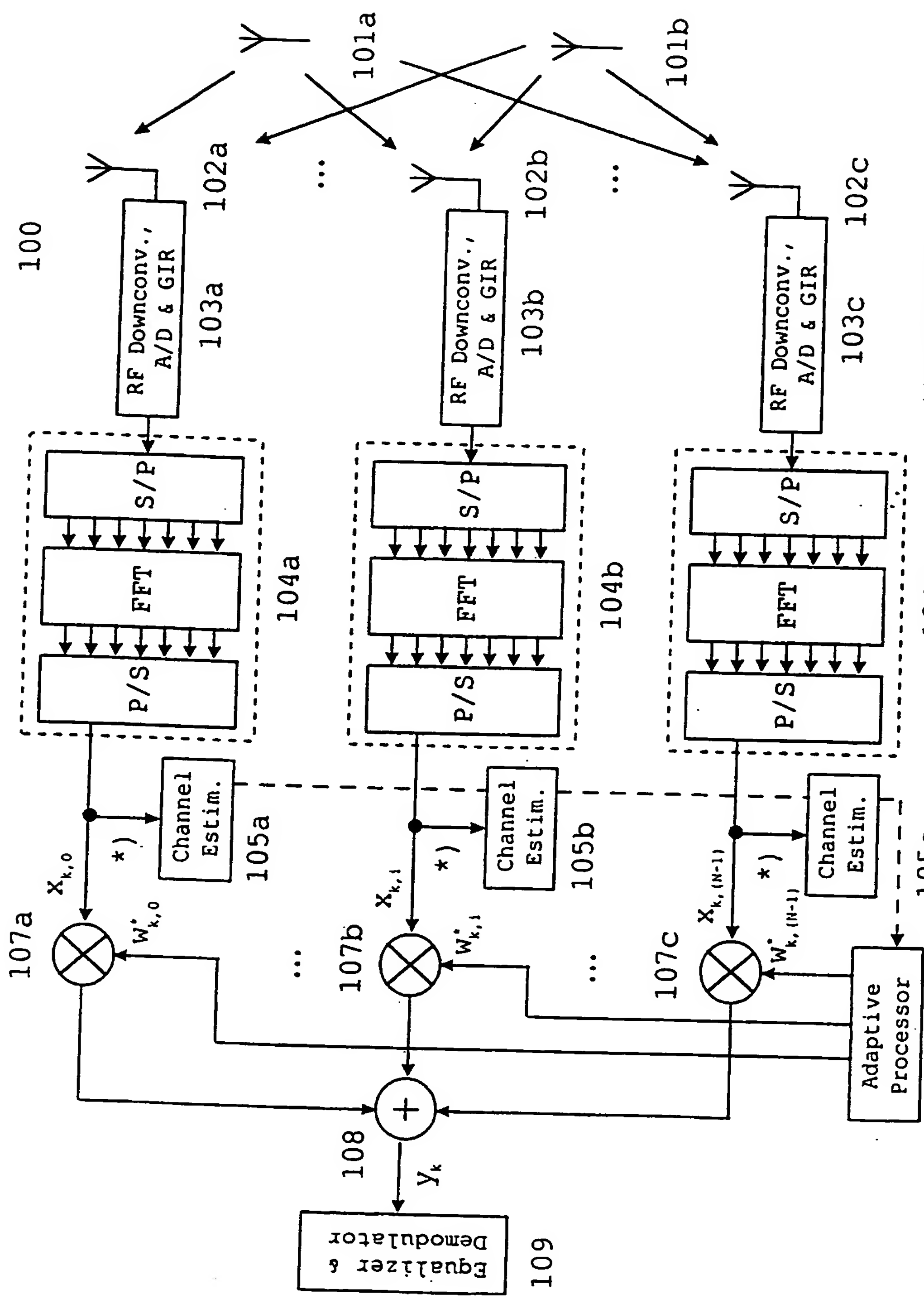
21. A method according to claim 20, comprising means for performing a load swapping, wherein the modulation scheme of subcarriers having a power level higher than a predefined threshold is increased, whereas the modulation scheme of subcarriers having a power level lower than a predefined threshold is decreased, thereby departing from the applied OFDM modulation scheme.

22. A method according to anyone of the claims 12 to 21, comprising at least one access point (AP) and at least two mobile terminals (MT), wherein orthogonalized antenna weighting vectors are applied to provide a duplex operation in said mobile transmission system in order to enable said access points (AP) to independently communicate with two or more of said mobile terminals (MT) at the same frequency and/or time.

23. A mobile telecommunications device,
characterized in that
it supports a mobile transmission system according to anyone of the claims 1 to 11.

24. A mobile telecommunications device,
characterized in that
it comprises a mobile transmitter and/or a mobile receiver designed for supporting a method according to anyone of the claims 12 to 22.

25. A computer program, which performs, when executed by a processor of a mobile telecommunication device, a method according to anyone of the claims 12 to 22.



*) Preamble Symbol

FIG. 1

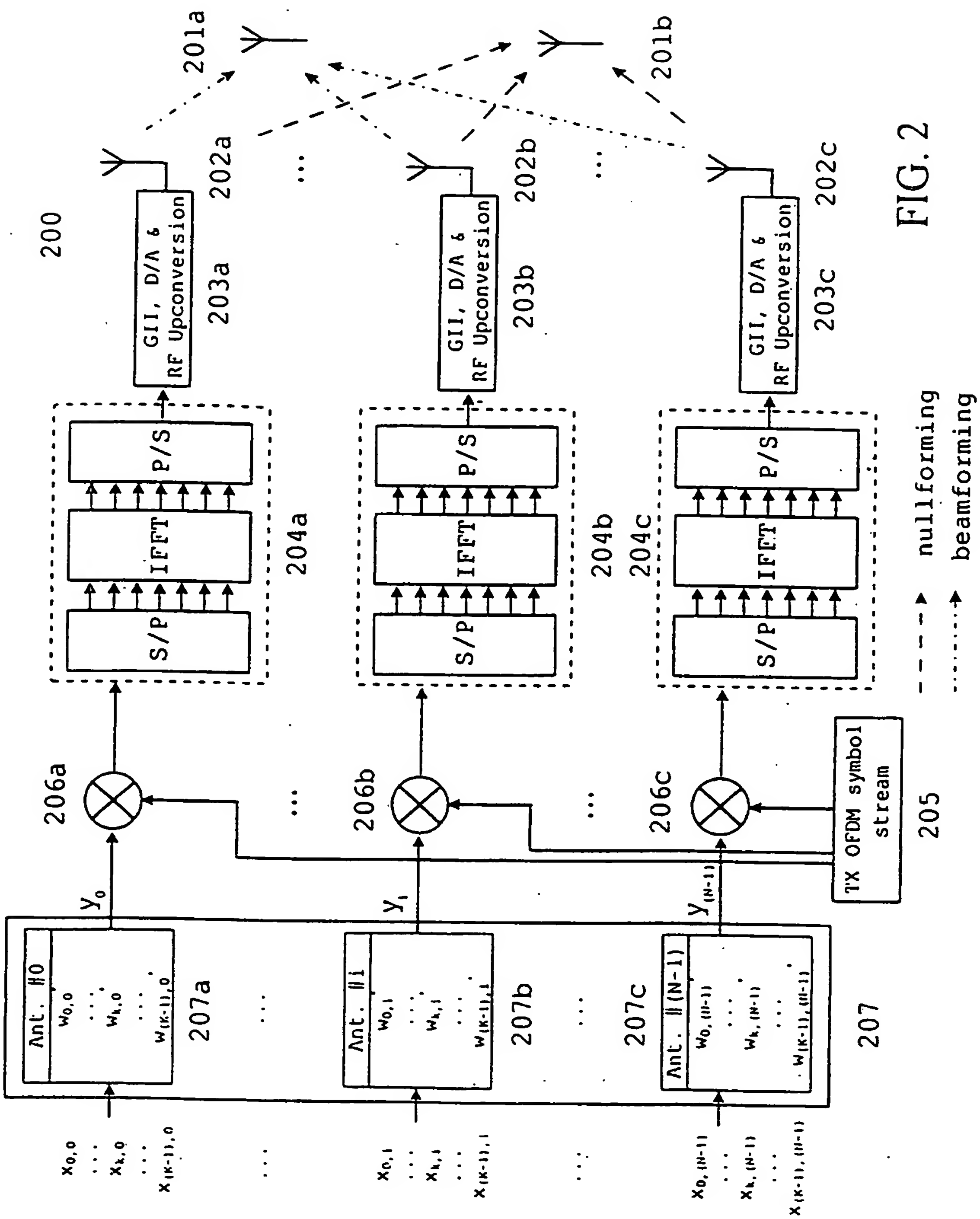

200
201a
201b
202a
202b
202c
GII, D/A & RF Upconversion
203a
GII, D/A & RF Upconversion
203b
GII, D/A & RF Upconversion
203c
P/S
IFFT
S/P
204a
P/S
IFFT
S/P
204b
P/S
IFFT
S/P
204c
206a
206b
206c
TX OFDM symbol stream
205
y_0
y_1
$y_{(N-1)}$
Ant. #0
$w_{0,0}$
$w_{k,0}$
$w_{(K-1),0}$
207a
Ant. #1
$w_{0,1}$
$w_{k,1}$
$w_{(K-1),1}$
207b
Ant. #(N-1)
$w_{0,(N-1)}$
$w_{k,(N-1)}$
$w_{(K-1),(N-1)}$
207c
207
$x_{0,0}$
$x_{k,0}$
$x_{(K-1),0}$
$x_{0,1}$
$x_{k,1}$
$x_{(K-1),1}$
$x_{0,(N-1)}$
$x_{k,(N-1)}$
$x_{(K-1),(N-1)}$
nullforming
beamforming

FIG. 2

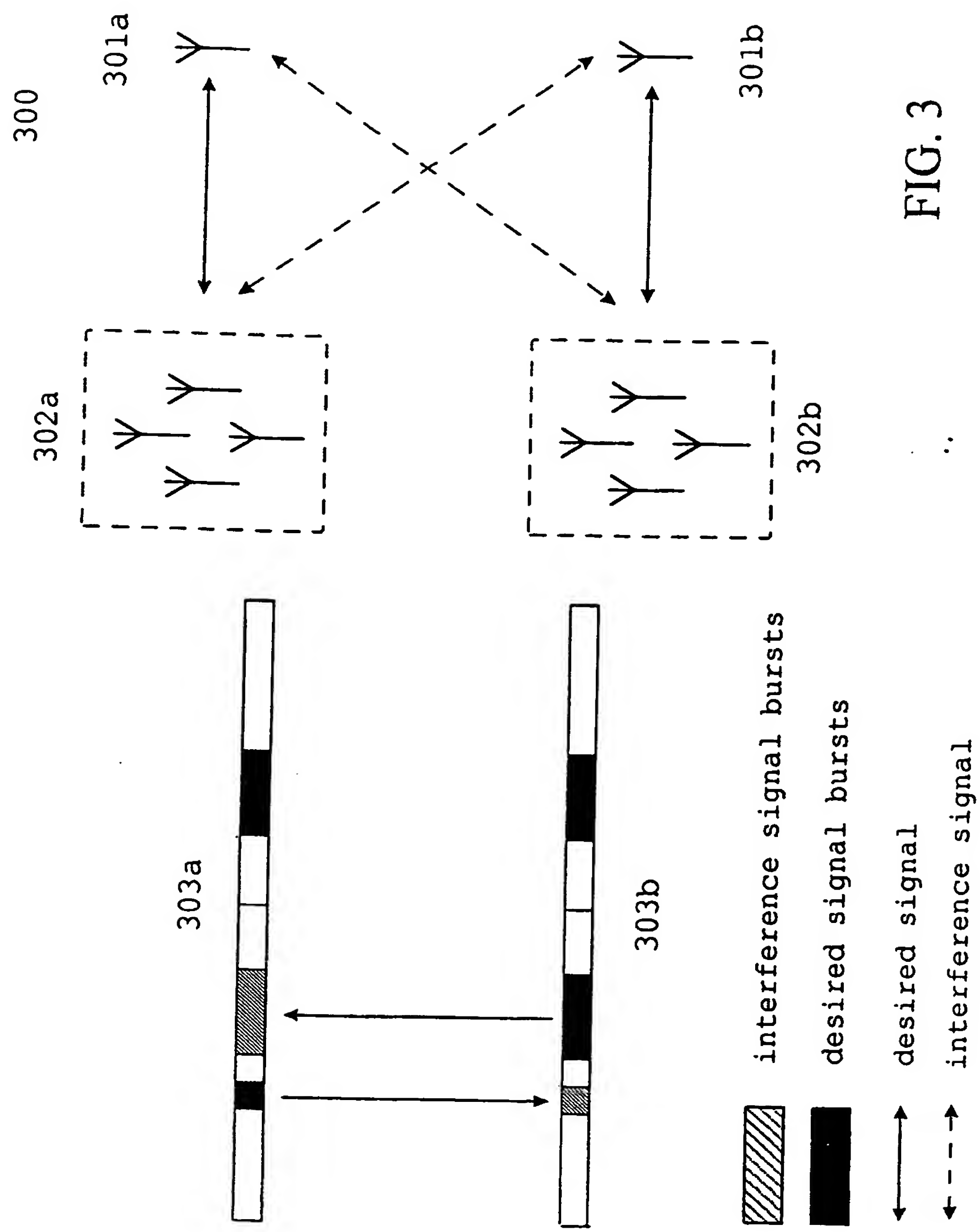

300
301a
301b
302a
302b
303a
303b
interference signal bursts
desired signal bursts
desired signal
interference signal

FIG. 3

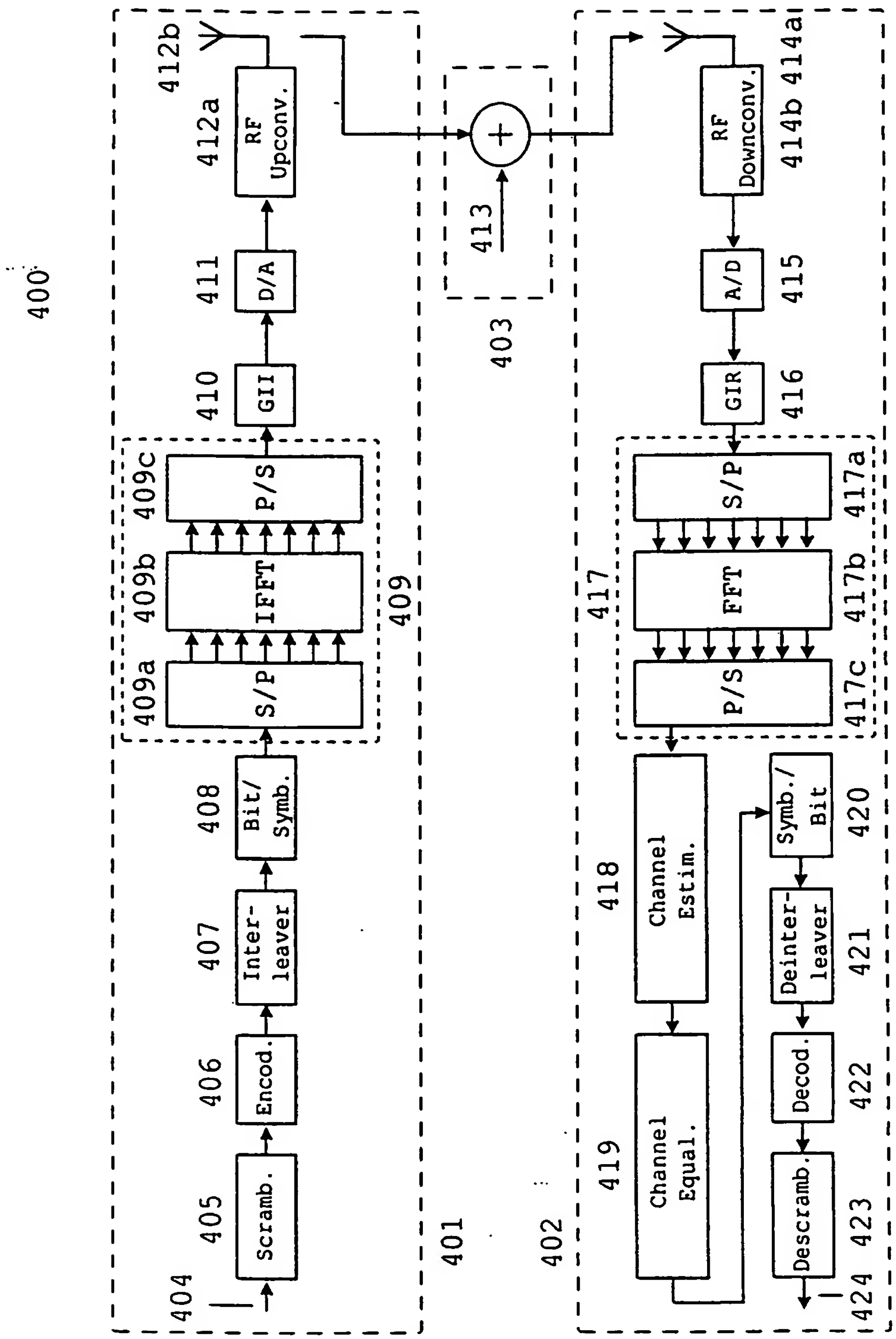

400
401
402
403
404
405 Scramb.
406 Encod.
407 Inter-leaver
408 Bit/Symb.
409
409a S/P
409b IFFT
409c P/S
410 GII
411 D/A
412a RF Upconv.
412b
413
414a
414b RF Downconv.
415 A/D
416 GIR
417
417a S/P
417b FFT
417c P/S
418 Channel Estim.
419 Channel Equal.
420 Symb./Bit
421 Deinter-leaver
422 Decod.
423 Descramb.
424

FIG. 4

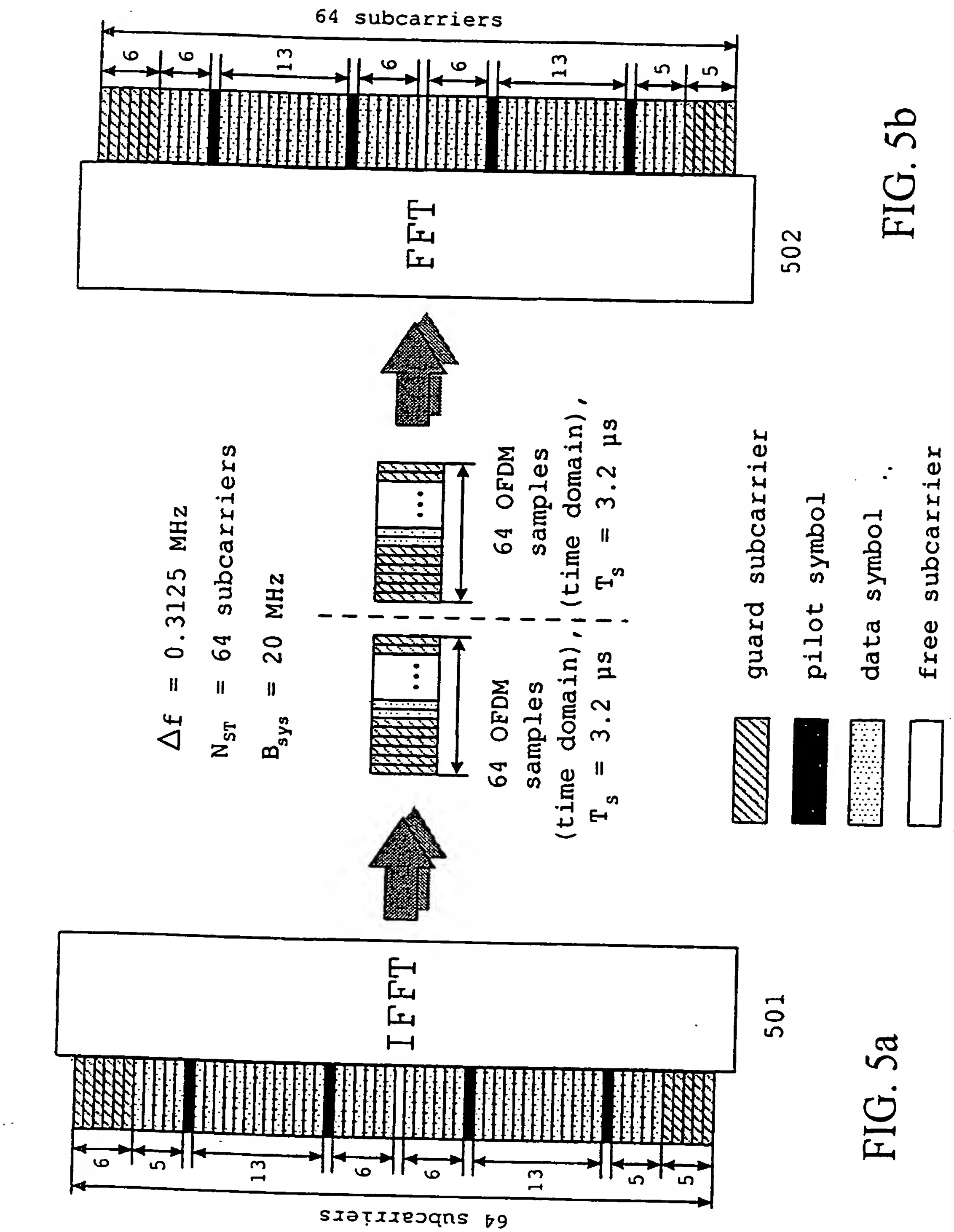

64 subcarriers
6
6
13
6
6
13
5
5
FFT
502
IFFT
501
6
5
13
6
6
13
5
5
64 subcarriers
Δf = 0.3125 MHz
N_{ST} = 64 subcarriers
B_{sys} = 20 MHz
64 OFDM samples (time domain), T_s = 3.2 µs
64 OFDM samples (time domain), T_s = 3.2 µs
guard subcarrier
pilot symbol
data symbol
free subcarrier

FIG. 5a

FIG. 5b

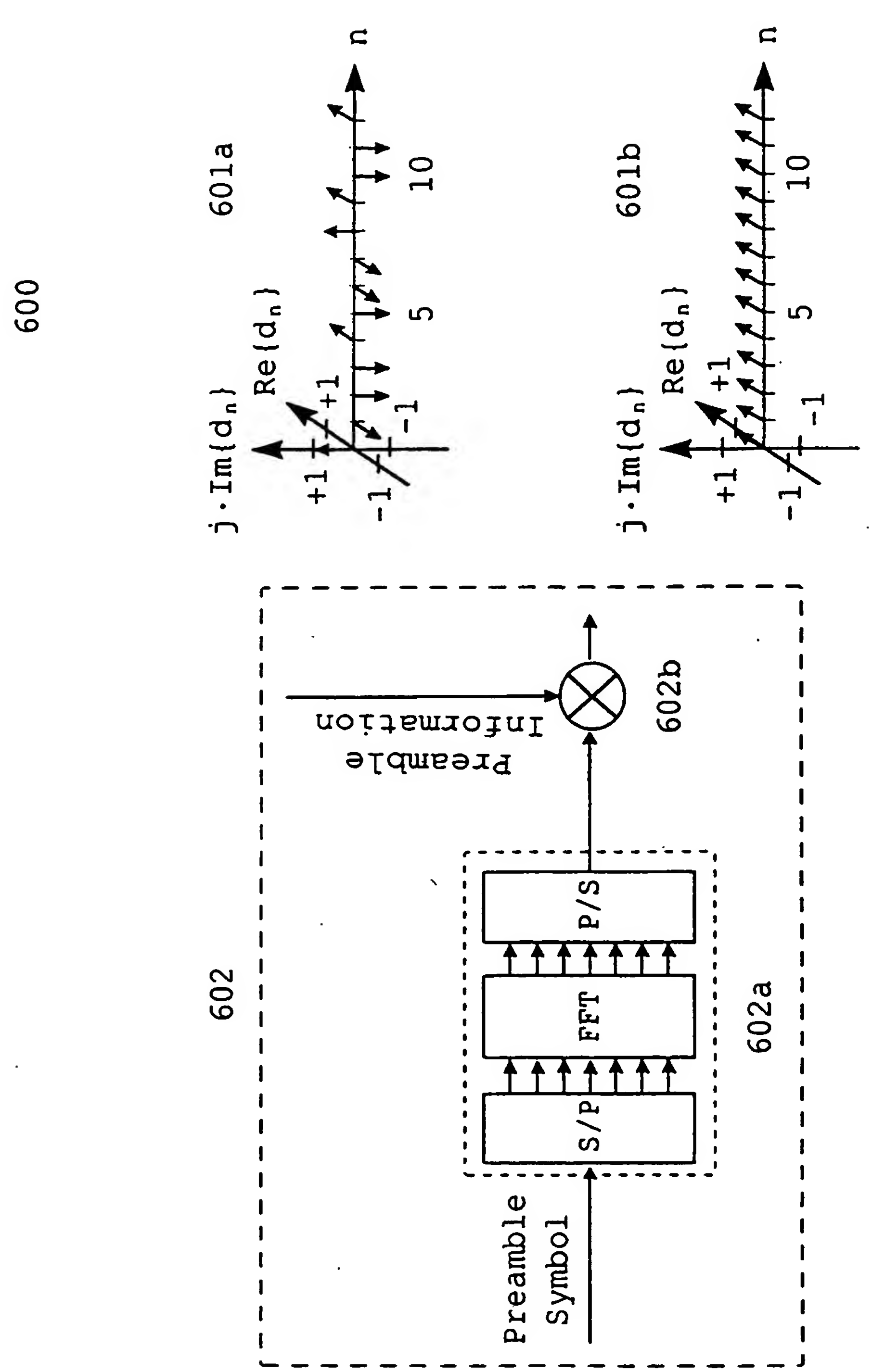
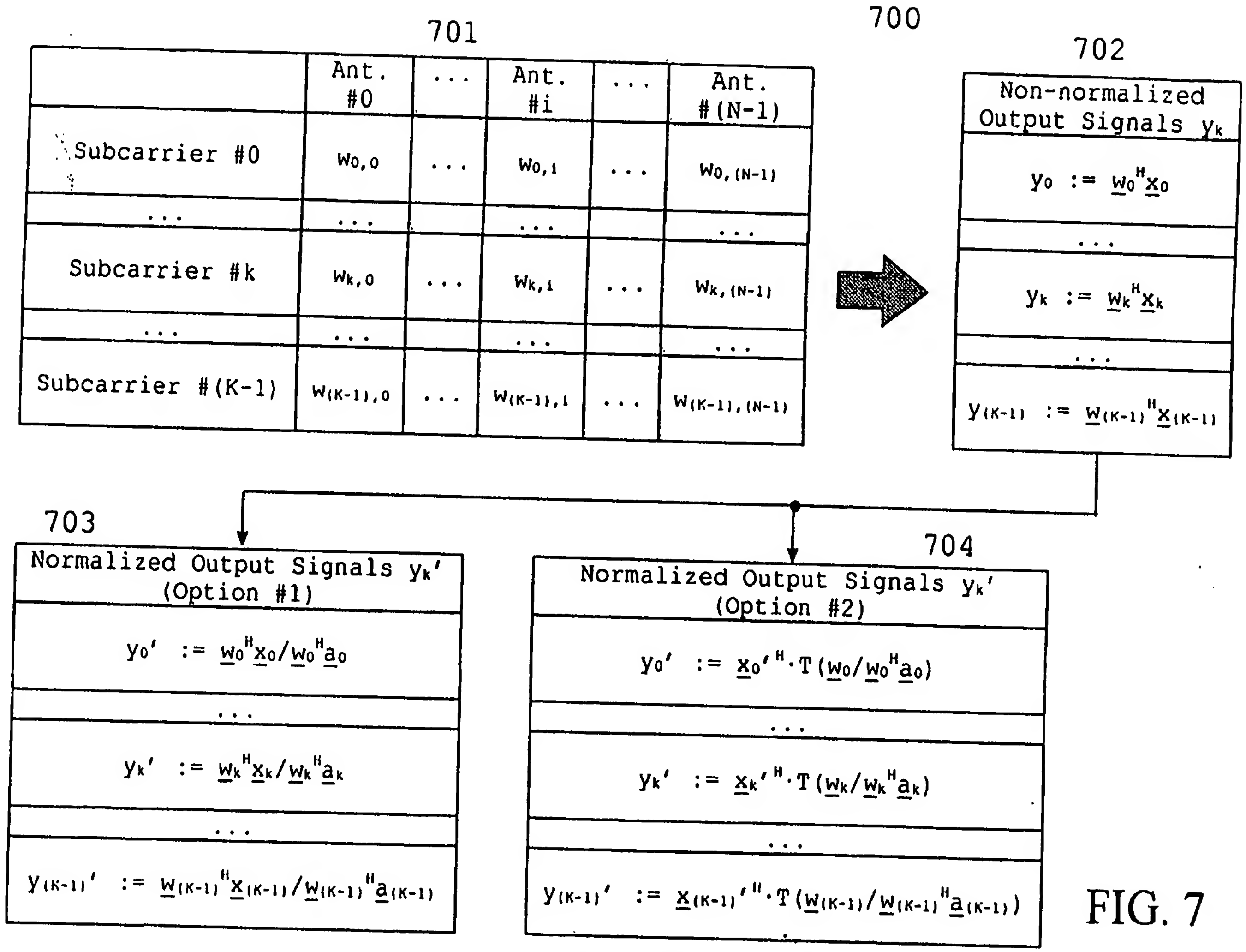

600
601a
601b
602
602a
602b
j·Im{dₙ}
Re{dₙ}
+1
-1
5
10
n
Preamble Symbol
S/P
FFT
P/S
Preamble Information

FIG. 6



$\underline{x}_k'$: swapped transmission signal
$T(.)$: threshold function

FIG. 7

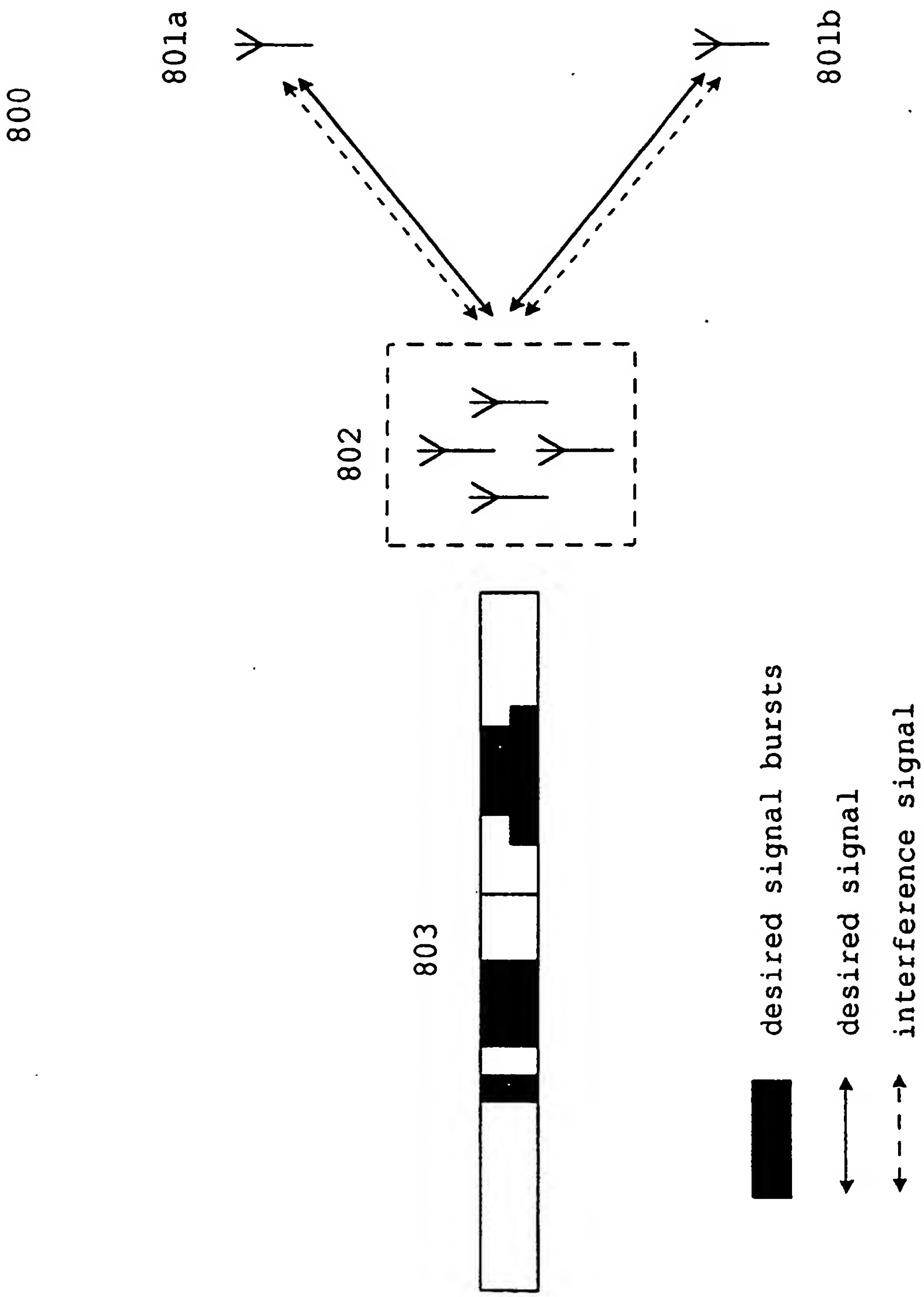

800
801a
801b
802
803
desired signal bursts
desired signal
interference signal

FIG. 8

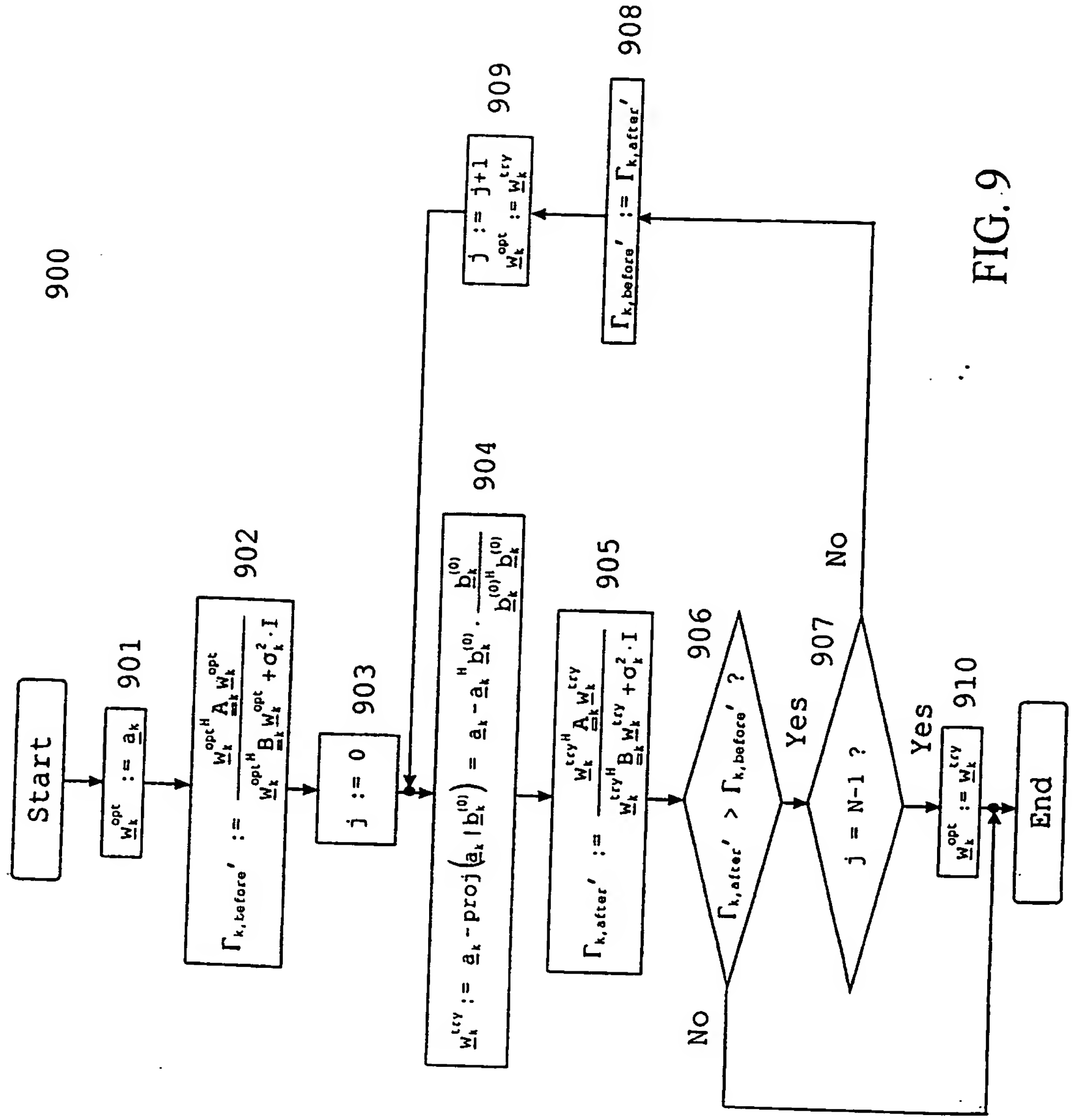

900
Start
901
$\underline{w}_k^{opt} := \underline{a}_k$
902
$\Gamma_{k,before}' := \frac{\underline{w}_k^{opt^H} \underline{\underline{A}}_k \underline{w}_k^{opt}}{\underline{w}_k^{opt^H} \underline{\underline{B}}_k \underline{w}_k^{opt} + \sigma_k^2 \cdot I}$
903
$j := 0$
904
$\underline{w}_k^{try} := \underline{a}_k - \mathrm{proj}(\underline{a}_k | \underline{b}_k^{(0)}) = \underline{a}_k - \underline{a}_k^H \underline{b}_k^{(0)} \cdot \frac{\underline{b}_k^{(0)}}{\underline{b}_k^{(0)^H} \underline{b}_k^{(0)}}$
905
$\Gamma_{k,after}' := \frac{\underline{w}_k^{try^H} \underline{\underline{A}}_k \underline{w}_k^{try}}{\underline{w}_k^{try^H} \underline{\underline{B}}_k \underline{w}_k^{try} + \sigma_k^2 \cdot I}$
906
$\Gamma_{k,after}' > \Gamma_{k,before}'$?
Yes
No
907
$j = N-1$?
Yes
No
908
$\Gamma_{k,before}' := \Gamma_{k,after}'$
909
$j := j+1$
$\underline{w}_k^{opt} := \underline{w}_k^{try}$
910
$\underline{w}_k^{opt} := \underline{w}_k^{try}$
End

FIG. 9

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 01 11 6931

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
A	US 5 999 800 A (YUN DONG UN ET AL) 7 December 1999 (1999-12-07) * abstract * * column 11, line 1 - column 13, line 43 *	1-25	H04B7/08
A	US 5 982 327 A (VOOK FREDERICK WERNER ET AL) 9 November 1999 (1999-11-09) * abstract * * column 7, line 15 - line 56 *	1-25	

TECHNICAL FIELDS SEARCHED (Int.Cl.7)
H04B

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	30 November 2001	Lazaridis, P

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
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L : document cited for other reasons
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EPO FORM 1503 03.82 (P04C01)

ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO.

EP 01 11 6931

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report. The members are as contained in the European Patent Office EDP file on
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30-11-2001

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
US 5999800	A	07-12-1999	KR	241503 B1	01-02-2000
			JP	10041733 A	13-02-1998
US 5982327	A	09-11-1999	NONE		

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For more details about this annex : see Official Journal of the European Patent Office, No. 12/82